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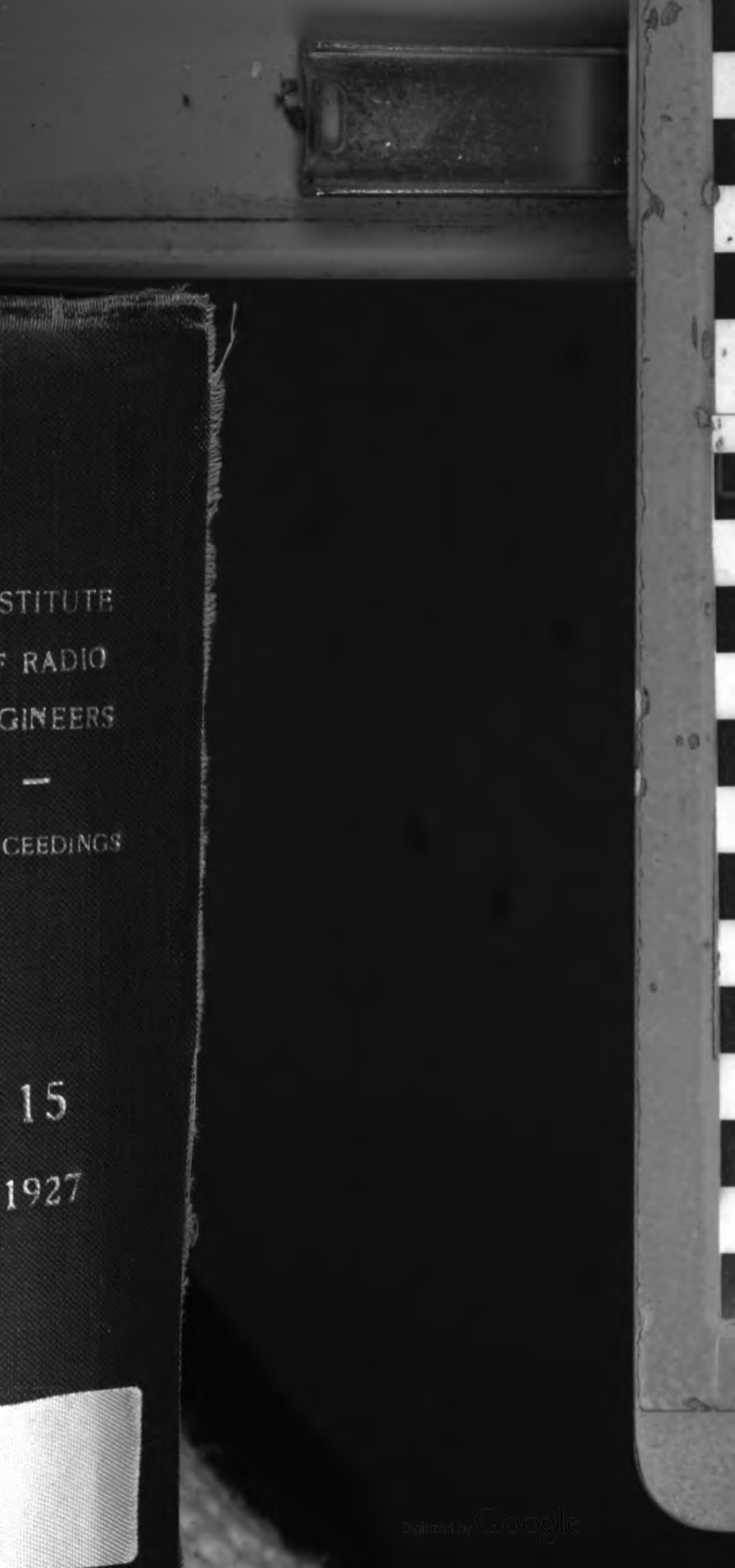

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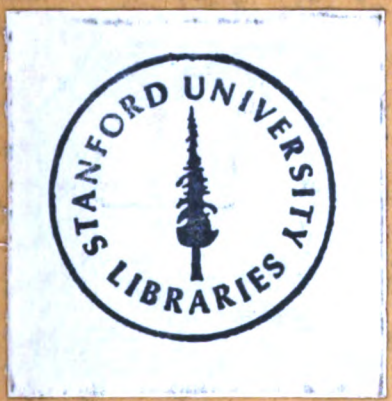

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INSTITUTE ACTIVITIES

New York Meeting, December 1.

At the December meeting of the Institute, held in the Engineering Societies Building, two papers were presented: one by Sylvan Harris, entitled "Notes on the Design of Resistance-Capacity Coupled Amplifiers," and one by M. Baumler entitled "Simultaneous Atmospheric Disturbances in Radio Telegraphy." The latter paper was read by Mr. D. K. Martin, and Mr. Harris' paper was read by Mr. D. E. Harnett. There was a very thorough discussion of the papers; participated in by Dr. J. H. Dellinger, C. M. Jansky, Jr., Roy A. Weagant, R. H. Marriott, Arthur Lynch and others. About two hundred members attended the meeting.

December Meeting of Board of Direction

At the meeting of the Board of Direction, held at Institute headquarters on December 1, the following were present: Donald McNicol, president; Dr. J. H. Dellinger, past president; W. F. Hubley, treasurer, and the following Managers: L. A. Hazeltine, Melville Eastham, J. V. L. Hogan, A. H. Grebe, R. H. Marriott and L. E. Whittemore.

The following were transferred to Member grade: Baldwin Guild, A. V. Simpson, L. E. Taufenbach, R. B. Stewart, H. W. Holcombe, D. K. Martin, George Uzmann and K. S. Johnson. Direct elections to Member grade: E. G. Widell and R. W. Seabury.

Fifty-seven associate members and seven junior members were elected.

Report was made of the establishment of a Section of the Institute at Hartford, Conn.

Membership Cards for 1927

The Card of Membership issued to Associates was first distributed to all Associates in the year 1926. The new cards for 1927 are ready and are being sent to all Associate members upon receipt of dues for the coming year. The new cards are of an attractive orange color.

Sections Committee

The Sections Committee, D. H. Gage Chairman, held a meeting at the chairman's office 253 Broadway, New York on the afternoon of November 23rd. The committee has done

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good work in connection with Section development and has plans under way looking to the establishment of additional Sections in territories where there is a large enough number of members of the Institute, or where such membership could be built up.

News of the Sections

Seattle Section

The Seattle Section held a meeting on November 6th in the Club Room, Telephone Building, Seattle, Washington. The meeting was addressed by Mr. Walter A. Kleist on the subject of "Picture Transmission as Employed by The Bell System." The paper was illustrated by lantern slides and was discussed by various members present including Mr. Libby, Roy Batch and C. E. Williams. Twenty-five members attended the meeting.

A meeting of the Section was held on December 4th at which Mr. Albert Kalin gave a talk on the subject of "Transformer Leakage."

Rochester Section

The Rochester Section held a meeting on the evening of November 19th. A paper was presented by Mr. H. A. Wheeler, of John Hopkins University on the subject "Applications of Thermionic Amplifiers". Fifty members were in attendance.

The Section held a meeting on the evening of November 30th at which Mr. W. A. McDonald, of the Hazeltine Corporation read a paper on the subject "Importance of Laboratory Measurements in Design of Radio Receivers."

Los Angeles Section

The Los Angeles Section held a meeting in the auditorium of the Y. M. C. A. of Los Angeles on the evening of September 20th. Mr. H. Pratt gave an interesting talk on the subject "The Problems of The Radio Engineer." Thirty-five members attended the meeting.

A meeting of the Section was held in the Commercial Club Rooms on the evening of November 22nd. A paper was

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presented by J. J. Jakosky on the subject "Phonographs Adapted to Radio Amplifiers" and a paper by H. M. Tremain on the subject "The New Electric Phonograph."

The meeting was preceded by a dinner. Thirty-five were in attendance.

PHILADELPHIA SECTION

The Philadelphia Section held a meeting on the evening of November 19th at which a paper was presented by Dr. D. Galen McCaa on the subject "Development of an Anti-Static Device". Thirty-five members attended the meeting. The next meeting of the Section will be held in the latter part of January, 1927.

CANADIAN SECTION

The Canadian Section held a meeting on the evening of November 24th in the Electrical Building of the University of Toronto. Professor A. M. Patience presented the paper "Output Characteristics of Amplifier Tubes" by Warner and Loughren. Twenty-one members attended the meeting.

The Section held a meeting on the evening of December 8th, at which Professor E. F. Burton of the University of Toronto presented a paper on the subject of "Physical Characteristics of the Human Voice."

WASHINGTON SECTION

The Washington Section held a meeting on the evening of November 10th in the Conference Room, Department of Commerce. A talk was given by Dr. A. H. Taylor on the subject "Recent High Frequency Progress". There were seventy-five members in attendance.

A meeting was held on the evening of December 8th in the rooms of Harvey's Restaurant of Washington. A paper was prepared by Dr. J. H. Dellinger on the subject "Effect of Fading on Broadcasting" was delivered by Dr. C. B. Jolliffe. There were forty members in attendance. The technical meetings of the Section are preceded by dinner at Harvey's Restaurant.

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Connecticut Valley Section

The Connecticut Valley Section held a meeting on the evening of December 3rd in the auditorium of the Hartford Electric Light Company. A paper was presented by Mr. Allen D. Cardwell on the subject of "The Engineering Design of High Voltage Variable Condensers." In addition there were motion picture exhibitions of interest to communication engineers.

ERRATA

The following corrections should be noted in the paper by Levin and Young in the Proceedings for October, 1926.

Page 681, 10th line from top, i should be I.

Page 683, equation 17, X should be γ: 13th line from top, $a=B_z$ should be $a=B\gamma$: 7th line from bottom, transmission should be transformation; 6th line from bottom $(a-1)$ should be $(Z-1)$.

Page 686 top line i should be I: 2nd line, $\frac{4i}{2i}$ should be $\frac{4I}{2I}$: 3rd line from bottom, $\frac{2}{7}$ should be $\frac{2}{9}$.

Page 688, last term of first equation should read $-j\frac{2\pi c}{\lambda}CV$

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PIEZO-ELECTRIC CRYSTAL-CONTROLLED TRANSMITTERS*

BY
A. CROSSLEY

(Naval Research Laboratory)

INTRODUCTION

The importance of producing vacuum tube transmitters which generate a constant frequency has never been seriously considered until recently, when the advent of broadcasting stations and the increasing number of ship and shore radio stations has demanded that such transmitters be made available.

The Naval Radio Service has been faced with this problem for a number of years particularly in the operation of the United States Fleet. Such a fleet generally consists of a group of vessels, numbering from 150 to 200 ships, which move as one unit and, for the greater part, their movements are controlled by radio. These vessels have one or more transmitters on board and are required to be in constant touch with each other and also with a group of shore stations. When we consider such a large number of stations, as represented by this Naval force it is very easy to imagine the confusion that will result if no means are employed to maintain a constant frequency for each station's transmitter.

Various means have been employed to hold constant the frequency of transmitters, but no absolutely satisfactory means has been devised which will maintain a constant frequency in vacuum tube transmitters which employ self oscillating circuits. This statement has special reference to transmitters which are required to operate over a band of frequencies and are dependent for plate and supply filament power on the usual ship's dynamo or shore stations power sources.

*Received by the Editor December 10, 1926.

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When we consider that the beat note of a continuous wave transmitter has to remain within a certain tolerance, say 350 cycles, it can be realized that the constant frequency condition becomes harder to meet as the frequency of the transmitter is increased. This can be readily seen when we consider that the frequency of a 4000-kc. transmitter cannot be changed more than 1/100 of one per cent before it has exceeded the specified frequency tolerance.

Realizing, after years of experimentation, that we could not meet the demands made on us by the Fleet with vacuum tube circuits employing the self-oscillating principle it was necessary to turn to some other means for meeting this demand. One means which has proved successful is the piezo-electric crystal-controlled transmitter. Such a transmitter has been found to meet all our requirements if suitable means are provided to keep the temperature of the crystal constant.

In order to cover development work undertaken at the Naval Research Laboratory the following paper is presented together with a resume of the art as observed.

The Piezo-Electrical Crystal

There are a number of crystalline substances such as quartz, tourmaline and Rochelle salts which have excellent piezo-electric and pyro-electric properties. All these are from an optical standpoint doubly refracting and possess asymmetric atomic structure. Bragg and Gibbs' show that alpha quartz which is piezo-electrically active has an unsymetrical hexagonal atomic structure, while beta quartz which has no piezo-electric properties is of a regular hexagonal atomic structure.

It is only natural to assume that any crystalline body which has double refracting properties and whose atomic structure is unsymetrical should be piezo-electrically active. F. Pockels[2] states that 20 out of 32 crystalline substances show some piezo-electric properties.

Considering the three commonly known piezo-electric crystals, i. e., quartz, tourmaline and Rochelle salts, we find that quartz is to be preferred. Rochelle salts although it has ten times the piezo-electric properties of quartz is not reliable. It is fragile, extremely hard to manufacture and

[1]Bragg & Gibbs—Proc. Royal Society, Vol. 109—1926.
[2]Winkelmanns Handbuch der Physik No. 2.

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its physical dimensions can be easily changed by handling, especially when subjected to contact with water. It also will not stand any electrical load; for instance, if used as a resonator in connection with the output of an oscillator of a few watts capacity it will break down. This breakdown will either be in the form of a series of mosaic cracks throughout the crystal or it will consist of a melting process wherein the crystal suddenly flattens out and assumes an isotropic state. If the power is increased the salts will return to a liquid state.

Repeated attempts have been made to make Rochelle salt crystals function as well as quartz, for controlling the frequency of a vacuum tube transmitter, but no success has been obtained in this endeavor. The Rochelle salt crystal is not mechanically strong enough to withstand the vigorous vibration which is met with in the quartz crystal when controlling the frequency of a vacuum tube transmitter. It is also possible that the hysteresis losses in the Rochelle salt crystal are such that they tend to damp out any properties of the crystal for generating a return piezo-electric voltage required for maintaining a vacuum tube circuit in an oscillating condition.

There is no literature available which shows the application of Rochelle salt crystals as a means for controlling the frequency of a vacuum tube circuit.

Tourmaline is too expensive to be considered as a commercial product and therefore resort has to be made to the use of quartz.

Quartz can be obtained in reasonable quantities in Brazil, Madagascar, Japan and the United States. Any quartz which has no flaws, intergrowths or optical twinning can be so manufactured that it has excellent piezo-electrical oscillating properties. By this we mean that such a crystal can be used to control the output of a vacuum tube oscillating circuit at one definite frequency and with maximum output.

Quartz will retain its physical dimensions if kept at a definite temperature. It will also stand considerable abuse, which accompanies its use in oscillation test circuits, where the crystal is heated momentarily to temperatures in excess of 45 deg. cent. and is subjected to frequent washing. Experience has demonstrated that crystals will hold the original oscillation frequency for periods in excess of ten

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months, when operated continuously in a high-frequency transmitter system. Other exacting tests have proved that quartz is the only material known which is satisfactory for use in crystal-controlled vacuum tube circuits.

The quartz crystal is hexagonal in shape and when in its true form has an apex at each end. The methods of mining and also the process of growing are such that the two apexes are rarely found on crystals which are purchased from the importers. In the majority of cases it is rare to obtain crystals having sides and one apex which are not chipped or cracked due to rough handling or poor mining methods.

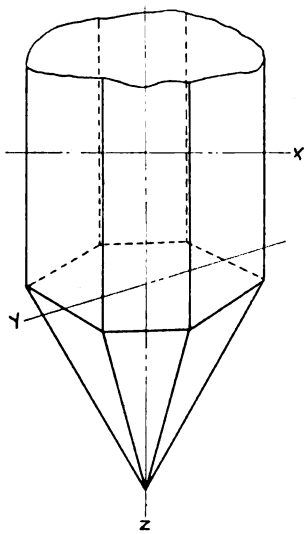

Figure 1

The usual crystal when received is similar in shape to that shown in Fig. 1. In this crystal the optical axis is parallel to an imaginary line *Z* which is drawn between two apexes. The electrical axes are of two types, one which is parallel to a line *X* drawn between the corners of the hexagonal sides and the other which is parallel to the line *Y* which is drawn between the opposite flat faces of the hexagonal sides. From this we note that there are three *X* electrical axes and three *Y* electrical axes and one optical axis. The optical axis is always at right angles or perpendicular to any of the electrical axes.

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Now cut a slab of quartz from the crystal as shown in Fig. 2 making this cut at right angles to the optical axis *Z*. Then in order to obtain a workable crystal we cut a slice from this slab as shown in Fig. 3. This slice is so cut from

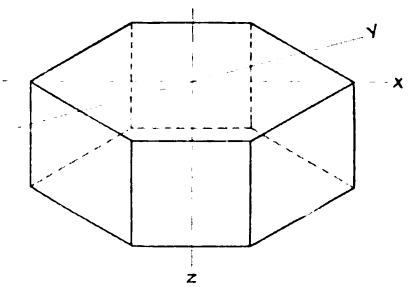

Figure 2

the slab that the slicing produces a crystal which sides are parallel to one of the *Y* electrical axes and at right angles to one of the *X* electrical axes. We now have a crystal whose thickness represents an *X* axis, whose length a *Y* axis and the depth a U or optical axis.

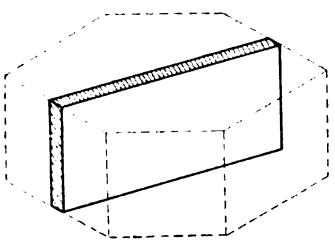

Figure 3

Methods of manufacturing the crystal from this point on to the perfect oscillating condition is being made the subject of another paper.

Having completed the cutting of the crystal which we will term the "Currie 3" or "zero angle cut" we find that there are three frequencies to which the crystal will resonate. One frequency corresponds to the *X* dimension, one to the *Y* dimension and the other to a frequency which is between *X* and *Y* axis frequency and is termed the coupling frequency. This coupling frequency depends on the dimensions of the *X* and *Y* axes. In round crystals as shown by Hund [4] the *X* dimension will produce 104.6 meters per mm.,

[4] A. Hund. Proc. I. R. E. August 1926.

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the *Y* dimension 110.5 meters per mm. and the coupling frequency is equal to 0.71 of the *Y* dimension wave length. In rectangular crystals the meters per mm. for the *X* dimension varies from 103.5 to 105.0 while for the *Y* dimension it varies from 110 to 117 meters per mm. The meters per mm. obtained for the coupling frequency cannot be stated because it depends on the dimensions of the rectangular form which may be square or any shape which the requirements demand. These figures are based on the true Curie cut and on crystals whose *Y* dimension is between 20 and 28 mm. If any cut is made which is at an angle from the Curie cut the meters per mm. will be greater, especially for the *X* dimension oscillation.

Rectangular crystals are to be preferred to round crystals, first because they are cheaper to make and second because they will control a greater radio frequency output without cracking or chipping. The latter condition is probably explained by the uneven stress conditions present in round crystals when they are oscillating under influence of radio frequency currents.

History

P. and J. Curie[2] first investigated quantitatively the piezo-electric properties of quartz and derived equations showing the relation between the applied pressure and the piezo-electric charge on the faces of the crystal. They also showed the converse effect where an electric charge on the crystal would produce a change in crystal dimensions.

Since this disclosure, various uses have been made of piezo-electric crystals, namely, as pressure gauges, loud speakers[5] and sound transmitters for underwater signalling[4].

Cady[7] first discovered that quartz could be employed as resonators and as such to be used as standards for precision frequency determinations. Cady later discovered that crystals could be used to hold the frequency of self-oscillating circuits constant and also that crystals could be made to control the frequency of a vacuum tube circuit. It is this feature of crystal control that we are most interested in and in order to present this subject we will consider each step made by various investigators in producing vacuum

[2]P. & J. Curie, Journal Physique, 2 series, Vol. 8. 1889—p. 149.
[5]A. Mc.L. Nicholson patent U. S. 1,495,429—May 1924.
[4]Paul Langevin British patent 145,691—July, 1921.
[7]Cady patent U. S. 1,450,246 Apr. 1923.

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tube circuits which would be capable of obtaining maximum radio frequency output from a quartz crystal.

Cady's work may be summarized by reference to Figs. 4, 5, and 6. The circuit shown in Fig. 4 is essentially a self oscillating vacuum tube circuit with a crystal placed across

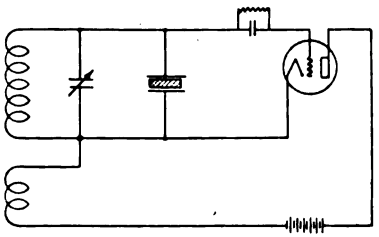

Figure 4

the grid tuning condenser. When the circuit is adjusted to the resonant frequency of the crystal there is a tendency in the crystal to keep the frequency of the circuit equal to that of the crystal. If the plate voltage, filament voltage and load remain the same, the crystal will hold constant the frequency of the circuit, but if one or more of the above conditions are changed the circuit will oscillate at any frequency which the circuit conditions permit.

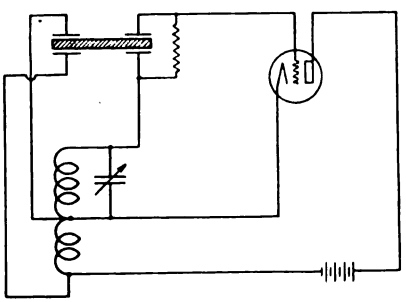

Figure 5

Fig. 5 is an elaboration of Fig. 4 and is employed to obtain a greater piezo-electric voltage for controlling the frequency of the circuit. The greater piezo-electric voltage is obtained by the use of the plate feed back principle represented by the extra set of plates on the crystal. The operation of this circuit is limited by the same conditions which are cited for the circuit shown in Fig. 4.

It has been our experience that any method which depends on any self-oscillating conditions in a vacuum tube

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circuit in addition to piezo-electric control is dangerous for two reasons: i. e., first, because of the danger of frequency shifting and second it is very easy to crack or chip crystals in such circuits. This latter case is exaggerated when we tie in the crystal oscillating circuit with an unbalanced amplifying system where the radio frequency current feedback from the amplifying circuit is sufficient to supply enough additional current through the crystal to cause it to heat up and crack.

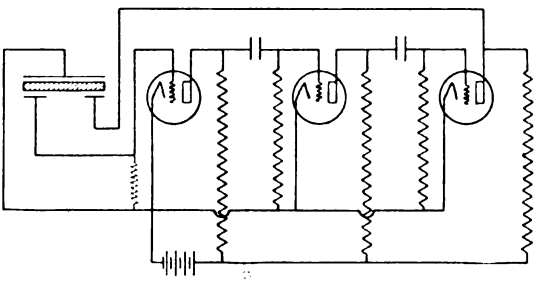

Figure 6

Fig. 6 is the first circuit known to the art wherein the crystal with the associated amplifying circuits comprise a system in which the crystal is the only control for the generator frequency. In this circuit the initial piezo-electric charge on the grid is amplified through three stages of resistance-coupled amplification, and by the means of a third contact plate on the crystal this amplified charge is applied to the crystal in the right phase relationship to reinforce the initial charge and by this process assist the circuit in generating radio frequency currents.

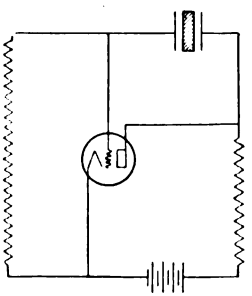

Figure 7

Pierce[5] later developed a circuit shown in Fig. 7 which is capable of generating by the use of crystal control a

[5]Pierce Journ. AMN. Academy of Arts and Sciences, Oct. 1923.

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source of constant frequency. In this figure the crystal is placed between grid and plate of a vacuum tube and a resistance load is inserted in the plate circuit. A grid leak is employed to hold the grid at a certain voltage with respect to the filament.

Fig. 8 is a modification of Fig. 7, wherein an inductance is substituted for the plate resistance. Both of the

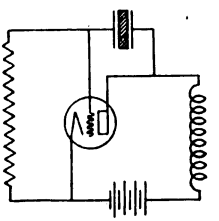

Figure 8

Pierce circuits function in the same manner as the old De Forest Ultraudion circuit. In the Ultraudion circuit, a tuned circuit is interposed between grid and plate as shown in Fig. 9 and a choke coil was employed as a load in the plate

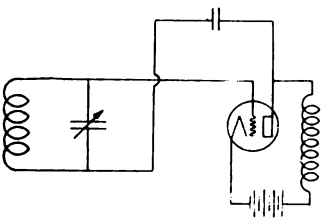

Figure 9

circuit. This choke coil was so constructed that it acted as a capacitive load for all frequencies to which the tuned circuit was resonant.

In the Pierce circuits the crystal functions as a tuned circuit having a preponderance of inductance while the plate load for the condition of oscillation has to be capacitive. To accomplish this end, Pierce uses a very large inductance coil in the plate circuit. The true inductance and the distributed capacity of the coil system used in this circuit has to be such that it will resonate to a lower frequency than that of the crystal before the circuit will oscillate. In the case where resistance is used, as in the plate circuit of

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Fig. 7, it is the distributed capacity of the resistance together with plate-filament capacity that affords the capacitive reactance required for the oscillation condition.

If the proper precautions are observed with the Pierce circuit with respect to the capacitive plate load condition, any crystal can be made to trigger off this circuit into the oscillating condition. In view of the fact that a grid leak is employed and the plate load is a resistance or a large inductance it is not possible to obtain the rated power output from a given tube with this circuit. This statement is based on the fact that the I^2R losses in the resistance and inductance are considerable and the grid leak method of biasing is inefficient for reasons which will be explained later in this paper. There is another objection to the Pierce circuit and that is the broad impedance curve of the plate circuit, which permits the generation of a number of oscillations at one time should the crystal be so constructed that there are two possible oscillations, for the X dimension or the Y dimension frequency may be very close to the coupling frequency with the result that both frequencies will be heard.

Piezo-Electric Crystal Research Work At the Naval Research Laboratory

Realizing the limitations of piezo-electric crystal circuits then known to the art, further development work was carried on at the Naval Research Laboratory to determine whether piezo-electric crystals could be employed to control any system which could permit a reasonable radio frequency output.

J. M. Miller formerly of this Laboratory developed the circuit shown in Fig. 10. This circuit is similar to Pierce's

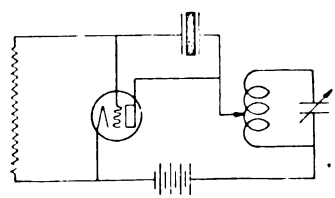

Figure 10

circuit with the exception that Miller employed a tuned plate circuit and a variable tap on the inductance. The tuned circuit permitted tuning to any desired frequency, thus ex-

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cluding undesired frequency oscillations. The variable plate tap permitted matching of tube impedance to circuit impedance whereby maximum power transfer was possible. Low loss inductance and condensers were employed thus reducing I^2R losses in the tuned circuit to a minimum

Miller also developed the circuit shown in Fig. 11, which circuit is the fundamental Navy circuit. In this circuit the crystal is placed between grid and filament instead of be-

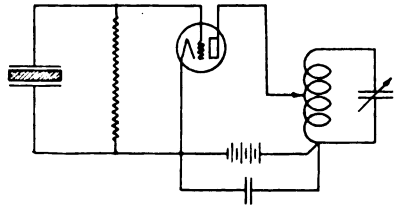

Figure 11

tween grid and plate as in Fig. 10. With this circuit the load for the plate circuit should be inductive in order that a condition for oscillation be obtained. The action of this circuit with reference to the oscillating condition is similar to the well known Hartley self-oscillating circuit. This may be better understood by stating that the crystal, being equivalent to an inductance, is similar to the grid coil of the Hartley circuit, while the inductive load in the plate circuit of the crystal oscillator is identical to the plate coil of the Hartley system.

Miller demonstrated that he could make both circuits oscillate with crystals of different frequency ratings. Preference was given to the use of the circuit shown in Fig. 11 because in such a circuit there is no tendency for short-circuiting the high-voltage plate circuit should the crystal crack or slide out from between the contact plates thus causing the plates to come in contact with each other. Experiments with high-frequency crystals show that good output is obtained with high as well as low-frequency crystals when employed in the circuit shown in Fig. 11.

The problem of obtaining greater output from the crystal oscillating circuit and the amplification of this output on low frequencies was assigned to A. Crossley. The Miller circuit Fig. 11 was used as a foundation and efforts were made to increase its radio frequency output.

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A study of this circuit showed that the crystal-controlling voltage was reduced materially by the fact that a grid leak was shunted across it. This grid leak provided a shunt path for the radio frequency piezo-electric control voltage and at the same time carried the rectified grid current required for obtaining the negative voltage for biasing the grid. Miller suggested the use of a choke coil in series with the grid leak to eliminate that part of the radio frequency voltage loss that is due to the direct flow of current through the grid leak resistance. This materially increased the output of the crystal oscillating system.

Crossley eliminated the second load on the crystal by substituting a battery for the grid leak, connecting the negative terminal of the battery to the low radio frequency potential side of the choke coil shown in Fig. 12.

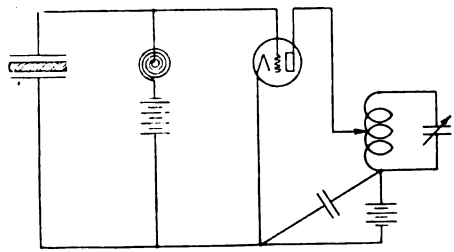

Figure 12

The object of the use of the battery was to keep the swing of the crystal-controlling voltage over on the negative side of the grid voltage-plate current characteristic curve, thus eliminating the load in the grid-filament path which is obtained when the grid swings positive past the zero grid voltage point on this curve. The act of swinging over the positive side of grid voltage-plate current characteristic curve causes a current flow between grid and filament which represents an I^2R loss. Any method for eliminating this grid current flow or rendering it of negligible value will also permit the crystal controlling voltage to be kept at a maximum and therefore permit a maximum output to be obtained from the circuit.

In actual practice, there is a small amount of grid current flowing, but the grid voltage swing is for the greater part over on the negative side of the grid voltage plate-current characteristic curve. This grid current flow is due principally to the fact that the grid battery employed does

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not block the plate current to zero, because a small amount of plate current is necessary before the crystal circuit will start oscillating.

The use of the battery increased the output of the circuit tremendously. Using a 600-kc. crystal it was possible to obtain one watt output with the original circuit employing the grid leak while with the choke and battery an output of 21 watts was obtained. The tube used in this experiment was the U V 210 type rated at 7½ watt allowable plate dissipation. The efficiency of this circuit was 65 per cent figuring plate input watts against radio frequency output. The plate voltage for this test was 650 volts, while the grid battery was 80 volts.

Further experiments with 50-watt tubes and also special tubes, have shown that as high as 100 watts output can be obtained when employing high-frequency crystals in the 3000 to 4000-kc. band. It is not possible to obtain such outputs when employing higher or lower frequency crystals. With the lower frequency crystals the feed-back afforded by the grid-plate capacity is proportional to the frequency, while the capacity between crystal plates is reduced as the frequency of the crystal is decreased. In other words the charge on the crystal is reduced as the frequency of the crystal is decreased and correspondingly the piezo-electric controlling voltage from the crystal is likewise reduced. To make up for the reduced charge on the crystal it is customary when using lower frequency crystals to increase the plate and grid battery voltage.

There is, however, as stated before, a certain high frequency where maximum output is obtainable and on either side of this frequency there is slow decrease in output. Around 3500 kc. is approximately the peak output frequency point, while at 12,000 and at 100 kc. are the frequencies where minimum output is obtained. These output ratings, are based on safe crystal current-carrying ratings, which was first shown by A. Hoyt Taylor. Taylor shows that for the different frequency crystals there is a safe working current at which the crystal can be operated and if this point is exceeded the crystal will heat up and crack.

This condition can also be tied down to a safe wattage dissipation in the crystal, but not knowing the resistance of the crystal we can only consider it from a current standpoint. In the 3000 to 4000 kc. band, electrostatic voltmeter

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and thermal ammeter readings in the crystal circuit show approximately five watts loss in the crystal. It is only at this and higher frequencies that an electrostatic voltmeter can be used to measure the voltage across the crystal, because the shunting effect of the capacity of the meter at the lower frequencies is such that it seriously reduces the output. When measurements were made with a 500-kc. crystal, the placing of an electrostatic voltmeter across the terminals of crystal holder was such as to reduce the output to less than one-half of the original output. This reduction in crystal piezo-electric controlling voltage is due to the fact that the plate-grid feed-back voltage is divided between the two capacities, i. e., the voltmeter capacity and the crystal capacity, while at the lower frequencies the crystal capacity is nearly equal to the voltmeter capacity.

This can be shown further by reference to Fig. 13 where Cr is the capacity equivalent to the crystal, capacity Cv

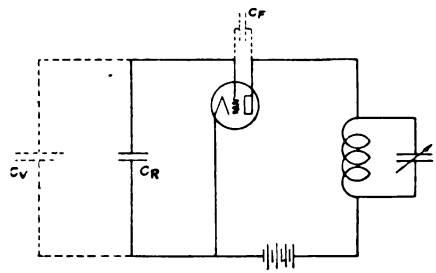

Figure 13

represents the voltmeter capacity of Cf the grid-plate feed-back capacity.

From the above figure we can note that it is not possible to obtain true voltmeter readings due to the presence of the shunt capacity Cv which divides the feed-back charging voltage into two paths, thus robbing the crystal of the maximum charging voltage from the plate circuit. Although we have cited the case of the shunt capacity afforded by the voltmeter, it is also true than any extraneous capacities, such as long leads from crystal to grid, poor design of crystal holder which permits additional capacity other than that of the crystal contact plates, and choke coils which have high distributed capacities, will also produce the same effect.

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The capacity between contact plates of crystals in the range between 100 and 12,000 kc. respectively, varies from 12 to 125 micromicrofarads. When we consider the capacity of the crystal and the fact that the grid-plate feed-back capacity is constant, it can be readily seen when using the same plate voltage, that the charge delivered to the crystal is reduced with the decrease of frequency. An example of this case is cited by comparing charges delivered to two crystals, i. e., a 500 and a 4000-kc. crystal when employed in a 7½ watt U V 210 tube circuit. With the 4000-kc. crystal the charge is 64 times as great as that delivered to the 500-kc. crystal. This increased charge on the crystal with the increase of frequency explains why it is possible to obtain greater radio frequency output at high frequencies and why we have to be so particular about low-frequency crystals and their associated circuits. With the low-frequency crystals it is imperative that we employ the right amount of grid biasing voltage for the condition of maximum output, while with the high-frequency crystals it is possible to eliminate the biasing voltage and still obtain good output. It is, of course, understood that the elimination of the biasing battery means a reduction in the efficiency of the circuit and sluggish oscillating action of the crystal, especially when we use the crystal as a master oscillator.

An improvement on the Miller circuit was made by Crossley and is shown in Fig. 14. This improvement pro-

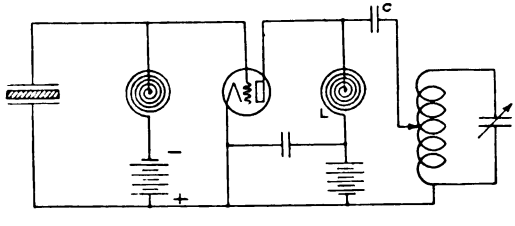

Figure 14

vides for the isolation of the radio frequency output circuit from the high-voltage direct-current circuit thus preventing the operator from accidentally coming in contact with the high-voltage direct-current supply. The insertion of the choke coil L and the condenser C in the circuit permits the segregation of the radio frequency and direct-current cir-

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cuits. It is good practice to make the resonant period of the plate choke coil L equal to a frequency which is lower than that of the crystal, thus making the choke coil a capacitive reactance at the crystal frequency.

Crystal Holders

The subject of crystal holders is very important. Experiments conducted by Crossley, particularly in the low-frequency range, show that the crystal will become inoperative if any dirt or moisture comes in contact with the crystal. If a crystal is placed in a circuit and started oscillating and a minute drop of water or oil is placed on the crystal, it will immediately stop oscillating. The stopping of oscillations may be explained when we consider that for best operation the top crystal contact plate is separated by a minute air cushion from the surface of the crystal when the crystal is oscillating and the introduction of moisture in place of the air causes a load to be placed on the crystal. This latter condition is similar to the use of mercury as a contact surface for the crystal, which type of contact adheres so closely to the crystal that it damps out any oscillation that tries to start up.

From the above facts it is imperative that the crystal be placed in a hermetically sealed container where no moisture or dirt can come into contact with it.

It is necessary that capacities other than that between the crystal contact plates be kept as small as possible, thus eliminating the charging losses occasioned by extraneous shunt capacities. For reliable operation and maximum output the crystal contact plates should be intimately touching the surface of the crystal. Lapped surfaces on these plates are to be preferred, while the weight of the upper plate should be kept to a minimum. No restriction of up and down movement of the upper plate should be tolerated. Light spring pressure can be applied to this plate but for best results no pressure other than the weight of the plate is necessary.

Retaining rings of bakelite or other insulating material or brass retaining pegs can be employed to hold the crystal in one fixed position with respect to the sides of the container. A holder[9] having all these features together

[9] Crossley patent U. S. No. 1,572,773—Feb. 1926.

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with means for restricting the tendency for the crystal to jump clear of the retaining pegs when being transported is shown in Fig. 15.

Experience has shown that any air-gap between upper surface of crystal and the contact plate means a great reduction in output and when used in regular power circuit the air-gap causes brushing between the surface of the

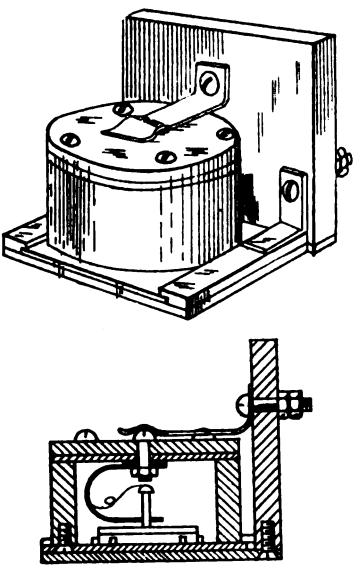

Figure 15

crystal and the plate, which in turn causes the crystal to heat and crack. Crystals which have been subjected to the brushing effect show a discoloration on the surface of the crystal at the place where the brushing occurred.

The frequency of any crystal changes with temperature and for absolute constancy of frequency it is necessary that some type of temperature control be applied either directly or indirectly to the crystal. One method is to place the crystal in a hermetically sealed container and by use of a thermostat and heating unit in this container maintain the crystal at a predetermined temperature. The second method is to place the crystal in a crystal holder of similar type to that shown in Fig. 15 and to secure this holder on a metal plate which can be maintained at a constant temperature.

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The heat from the plate will be conducted through the lower crystal contact plate direct to the crystal.

The metal heating plate can be kept at a constant temperature by circulating water through it, or a sub-compartment with suitable heating unit and thermostat can be attached to this plate. A thermostat can be employed with the water circulating system to turn on or off the current in a heating coil which is placed in the water intake line to the plate. This latter water-cooling method was developed by Taylor and applied by E. L. White to the high-power high-frequency transmitters at this Laboratory.

The importance of constant temperature control is appreciated when operating high-frequency crystals, as a change of 10 deg. cent. will change the frequency as much as one kilocycle in the 4000-kc. range. Extreme changes in temperature met with on board Naval vessels when cruising can change the crystal frequency as much as three kilocycles, which charge is very detrimental to perfect communication conditions. A remedy for this is to provide a thermostatic control which will maintain the crystal temperature above that which is ever encountered, throughout the year. This is identical with the practice now in force with reference to our Navy Standard 25-kc. crystal calibrator, which is used as a standard of frequency for the Navy.

Recent data on temperature coefficient in quartz crystals obtained by the Naval Research Laboratory show for the X axis there is a frequency change of 25 parts in a million per degree centigrade while with the Y axis a change of 50 parts in a million is noted. These data on electrical characteristics appear to show that the temperature coefficient of the elastic constant along the Y axis is approximately double that of the X axis, as we can assume that the frequency change must tie in with the mechanical change. This statement is based on our previous experiments on the resonant condition, in crystals, namely, the relation of meters per millimeter for respective electrical axes.

Several types of multiple crystal holders have been developed. One type employs the holder shown in Fig. 15 which was placed on a circular disk with a knob and pointer on the front of a panel for rotating the disk. Two contactors are placed behind the panel for making contact with each crystal holder as it was rotated past the indicated point.

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Crystal Controlled Power Amplifiers

Having obtained reliable operating conditions with the crystal controlled oscillator our next and most important problem was to amplify this output.

The first attempt to amplify the crystal oscillator output was made by resort to two stages of amplification. The first stage consisted of a 7½-watt tube while in the second stage two 50-watt tubes were employed. Grid leaks were used to bias the grids of the amplifier tubes. An output of 96 watts was obtained with this power amplifier at a frequency of 600 kc. Another stage of amplification consisting of three 250-watt tubes was added to this amplifier and the maximum output obtained was approximately 700 watts.

About this time Dr. Taylor and L. C. Young demonstrated that amplification of power from a 3000-kc. crystal oscillator was possible.

Considerable trouble was experienced from self-oscillations in the amplifier system. Various methods were employed to eliminate these undesired frequencies with no satisfactory result. Crossley discovered that these oscillations were in general of a high-frequency nature and that the only method of eliminating them was to place a resistance of a certain value in the plate lead, preferably as close to the plate as possible. The location of this resistance in the plate lead placed a load on the high-frequency circuit and if sufficient resistance was inserted the load would be too great to permit the grid-plate feed-back to cause a condition of self oscillation. The maximum value of the resistance did not exceed 300 ohms at 300 kc. and can be very low on sets of very high frequency; that is, small enough to have a negligible effect on the output of the amplifier circuit at the amplified frequency. This can be better understood when we consider that the plate circuit impedance at the desired frequency is at least 5000 ohms at medium frequencies for all types of tubes other than the one-kw. type, and the additional resistance placed in this circuit for purpose of stopping self oscillations never is greater than 6 per cent of the total circuit impedance. The impedance of the plate circuit which it is resonant to the self oscillation frequency is low, and the resistance referred to above is so located that it is in series with this resonant circuit.

The importance of shielding each stage of amplification was soon apparent and this was accomplished by providing

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metal containers for each stage. It was noted that there was sufficient feed-back through the grid plate capacity of the amplifier tubes to prevent maximum power amplification per stage. L. A. Gebhard and Miller suggested and demonstrated the power amplification gain which is possible when neutralizing the feedback and also applying a high value of biasing voltage to the grid of the amplifying tubes.

The neutralizing of the feed-back in the amplifier tubes permitted maximum grid excitation while the use of the high value of grid biasing voltage reduced grid-filament circuit losses to a minimum. A power amplification of 80 is obtainable by the method proposed above. A concrete case of this condition is cited when we amplified the output of a 7½-watt tube circuit by use of a UV 851 one-kw. tube and obtained 600 watts. From the above it can be noted that it is possible to reduce the number of stages of amplification very materially, thus eliminating troubles experienced with the excessive number of stages which are required when employing the old method of cascade amplification.

One more source of trouble was experienced in the development of the amplification system, and that was in the choke coils. It was found during numerous incidences that the choke coils would burn up, particularly the plate choke coils. These choke coils were of the single-layer and the pancake universal-wound type. An investigation of the reason for this burning effect showed that the burning occurred at frequencies which were close to the second, fourth, sixth, and other even harmonic frequencies of the fundamental of the choke coil. It then became necessary to make our choke coils such that the danger or burning frequencies would be other than that of the operating frequency of the transmitter.

If the transmitter is required to cover a broad band of frequencies a radical change has to be made in the choke coils. This change consists of using at least three choke coils, preferably of the universal-wound type, in a series connection. Each coil should have the same number of turns and the same shape and arranged on a bakelite or pyrex rod or form in such a way that the magnetic fields add. This multiple choke arrangement provides a method of obtaining in a concentrated form a choke coil which has nearly twice the inductance and two thirds the distributed capacity of the

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best type single-layer choke coil. It also has, by virtue of the addition of impedance of the respective coils, a high value of impedance at the dangerous harmonic frequencies. This latter characteristic makes this type of choke well suited for use in transmitters which employ the principle of frequency doubling or tripling to obtain super high-frequency oscillations. A choke coil similar to the type referred to above is shown in Fig. 16.

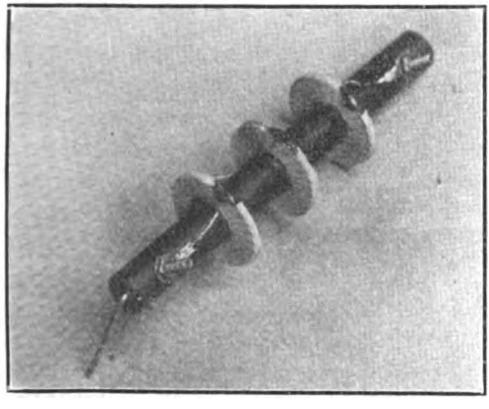

Figure 16

Having solved the major problems involved in the crystal controlled transmitter with reference to output the next problem was that of keying the system. Various methods were tried and abandoned due partly to sluggishness of action or the fact that too large a load was taken from supply source which furnishes the necessary negative voltage for blocking the grid of the control tube. A satisfactory system was finally obtained and may be explained by reference to Fig. 17.

In this figure there is shown one of the stages of amplification which it is intended to key. The grid circuit consists of the blocking condenser 1, choke coil 2, relay 4, high resistance 3 and the battery 5. Associated with the relay 4 is the key 7 and relay battery 8. The plate circuit is of the conventional type and consists of blocking condenser 9, antenna or dummy circuit 10 with the usual plate choke 11 and plate potential source 12.

The keying is accomplished by changing the grid biasing voltage from an operating voltage to a high blocking voltage

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through the agency of the relay 4 and the associated circuits. With the key 7 closed, current flows through the relay 4 and closes the contacts, thus permitting the grid lead to be connected to the low-voltage tap on the battery 5. The high resistance 3 is then placed across the remainder of

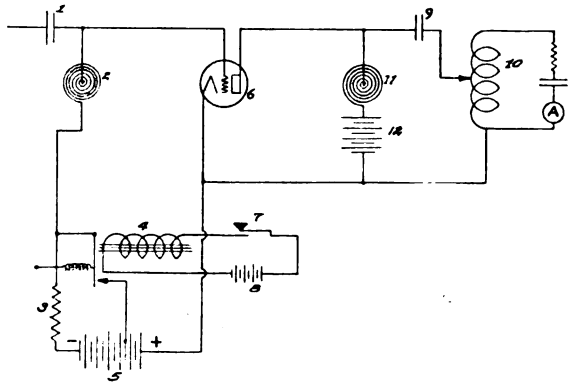

Figure 17

the battery and, due to its high value, it takes no appreciable load from the battery 5. When the key is up or in the open position the contacts on relay 4 spring back and disconnect the low voltage tap on the battery 5, thus through the resistance 3 making a path for the high blocking voltage to be impressed on the grid. The fact that there is no current flow through the grid circuit also indicates that there is no $I\ R$ drop over the resistance 3, thus we can, by means of the relay, change the grid voltage from an optimum operating value to an absolute blocking value. The use of the high-resistance also cuts down to a minimum the sparking and sticking of the contacts on the relay. With this method of keying it is possible to key at speeds in excess of 100 words per minute.

Complete Transmitter

Further work on this problem of amplification at frequencies from 150 to 600 kc. proved that it was possible to obtain an output of 13 kw. into a dummy antenna system. The complete details of the system which is capable of delivering this radio frequency power output may be explained by reference to Fig. 18.

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In this figure three stages of amplification are shown. The first stage consists of a 50-watt impedance coupled amplifier which feeds into a 1-kw. tuned amplifier stage and from this stage to a 20-kw. amplifier circuit.

The crystal controlled oscillator consists of a multiple crystal holder with the associated grid circuit which comprises the grid radio frequency choke and the source of biasing voltage. The plate circuit of this oscillator employs the parallel feeds through agency of the multiple choke coil, a source of high direct voltage, a 0.004 radio frequency by-pass condenser and a resonant circuit consisting of an inductance and two condensers with suitable radio frequency ammeter. Two condensers, one of which is variable are employed to permit tuning to resonance with the inductance over a given range of frequencies.

Voltage required for exciting the grid of the first amplifier tube is obtained from a tap on the inductance of the resonant circuit. Proper biasing voltage for the grid of the amplifier tube is obtained from a potentiometer and flows through the inductance of the resonant circuit direct to the grid. This method of biasing the grid of the amplifier tube eliminates the usual grid condenser and choke coil system.

The plate circuit of the first amplifier tube comprises a multiple radio frequency choke coil with the usual high voltage direct-current source and a resistance load circuit. The radio frequency output of the choke coil is delivered to the resistance load through the two by-pass condensers shown in the diagram.

The choke coil system is so constructed that it is resonant to a lower frequency than that of the range of the transmitter and therefore provides a capacitive load. This capacitive load prevents any tendency for feedback in the amplifier, thus saving the crystal circuit from any surge effects which tend to overload and break the crystal.

A variable contactor is used on the load resistance for obtaining optimum controlling voltage for the grid of the second amplifier tube. A 0.004-μf. by-pass condenser segregated the radio frequency resistance bad circuit from the grid biasing system. It will be noted that the grid biasing system of this tube includes the keying system referred to previously in this paper.

The plate circuit of the second amplifier has the usual parallel radio frequency feed circuit and in addition there

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is the balance or neutralizing circuit which comprises the counter-inductance and a 0.0002-μf. variable air condenser. This balance or neutralizing system is found to be very reliable and easy to adjust.

The third amplifier stage is identical to the second stage with the exception of the addition of the antenna load system. As will be noticed, the voltage for feeding the antenna is obtained from the drop across a condenser which is placed in series with the plate resonant circuit. This condenser was a 0.25-μf mica condenser with five taps of 0.05μf each. This condenser in addition to supplying required voltage for the antenna system also functions by virtue of its low impedance in reducing harmonic frequencies to a low value.

Protective Devices

Various kinds of protective devices were employed in this transmitter. The first and most important protective device was tied in with the biasing battery circuit and functioned in opening the filament and 2500-volt supply when no current was being supplied from the biasing battery.

The second device consisted of a water-flow protective relay circuit. When there was no flow of water through the water-cooled tubes the device opened the filament supply circuit and also the field of the high-voltage generator.

A circuit breaker was placed in series with the negative terminal of the high-voltage generator and when an overload was placed on the generator it opened the generator field circuit.

As a safety-first precaution, a condenser and a resistance was placed across the coil of the circuit breaker. These units maintained the negative side of the generator at ground potential should the breaker coil accidentally open. They also had a tendency to reduce line surges to a great extent by acting as a damping means.

Miscellaneous

Very small output is obtainable from a crystal on the lower frequencies required of the transmitter and for this reason a small amount of regeneration was employed in the crystal circuit. The regenerative feature is shown in Fig. 18, by the grid feed-back coil in the crystal oscillator circuit. On the higher frequencies this coil was short-circuited, for

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as previously stated in the paper, regeneration, especially at higher frequencies than 400 kc. is liable to crack the crystal.

The 50-watt stage of amplification was required for the low-frequency amplification, but can be dispensed with

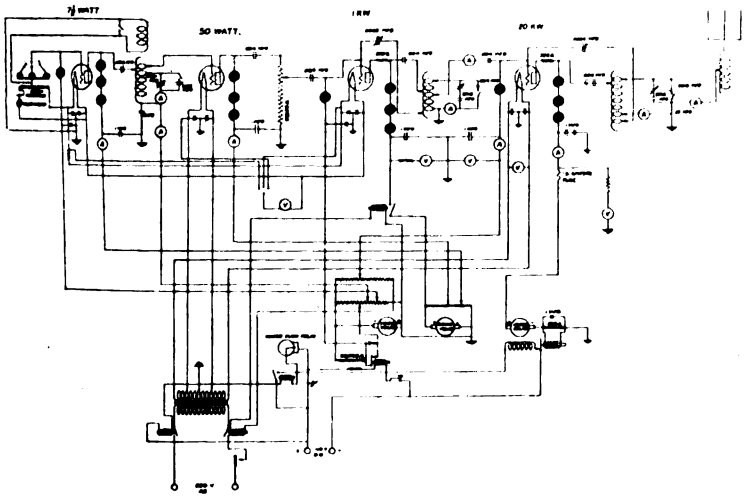

Figure 18

at the high frequencies. With well made crystals and proper circuits it is possible to use only two stages of amplification and obtain outputs in excess of 10 kw. when operating the transmitter over the range from 400 to 600 kc.

There was no need for frequency doubling in this transmitter and consequently no mention was made of it. Frequency doubling and tripling circuits were developed by Taylor and have been used very extensively in our numerous high-frequency transmitters which are in operation at N. K. F. This principle is also used in transmitters furnished by the Naval Research Laboratory to the Coast Guard, Army, Naval ships and stations.

The work covered in the construction of the low-frequency transmitter was undertaken by A. Crossley with the assistance of W. F. McBride. Messrs. Gebhard, Young, White and Taylor were responsible for the development of the high-power high frequency transmitters.

The development of the low power high frequency transmitter which derives its source of plate potential from alternating-current circuits was undertaken by R. B. Meyer.

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Meyer developed the crystal oscillator circuit which employs one crystal and two tubes with the split transformer plate supply circuit. Best results are obtained with such a circuit when the frequency of the supply source is 500 or more

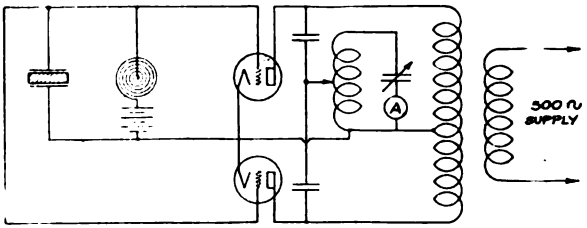

Figure 19

cycles. The Meyer circuit is shown in Fig. 19. This circuit is used in transmitters designed and built for the Army and the Marine Corps. A schematic wiring diagram of this transmitter is shown in Fig. 20. An inspection of this

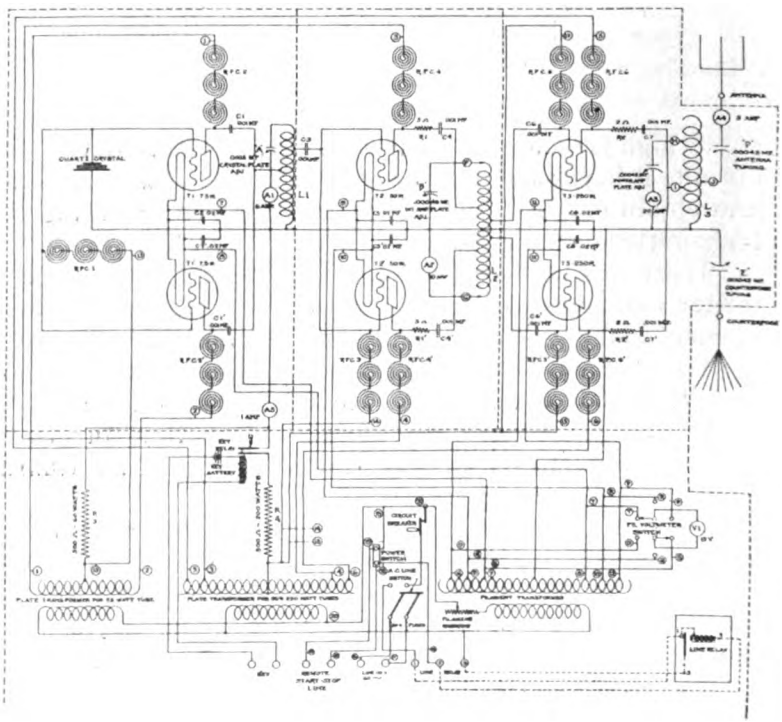

Figure 20

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diagram will show the automatic balance which is obtainable in the amplifier circuits when resort is made to the use of an alternating plate current supply. This balance is obtained by using approximately the same number of plate turns in each amplifier tube curcuit.

This transmitter is designed for frequency doubling and is capable of covering a frequency range from 3500 to 9000 kc. The rated output of the transmitter is 500 watts.

Figs. 21 and 22 are photographs of this transmitter showing front and side views.

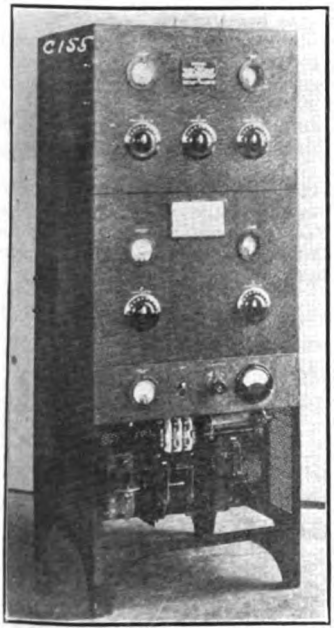

Figure 21

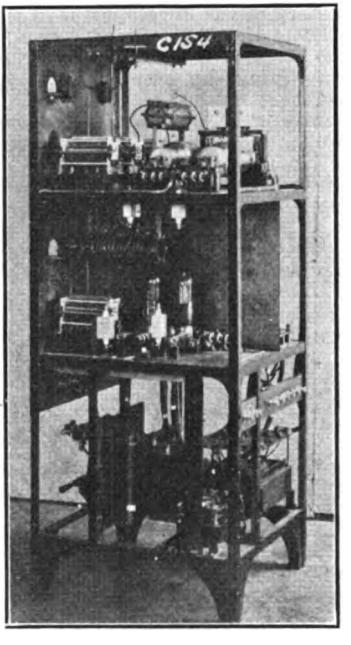

Figure 22

A method for obtaining more piezo electric controlling voltage was developed by Taylor[10]. This is accomplished by employing two crystals which have identical frequency characteristics and connecting these crystals in parallel with each other in the conventional circuit. Series stabilizing choke coils are placed in each crystal circuit for the purpose of holding both crystals in synchronism should temperature effects tend to change the natural frequency of the respective crystals.

[10]Taylor U. S. Patent No. 1,581,701—April, 1926.

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Taylor also developed a method[11] for obtaining three-phase source of radio frequency by employing three synchronized crystal circuits which feed into a *Y*-connected output circuit.

Among other developments of the Naval Research Laboratory is a means[12] for obtaining a crystal-controlled oscillating circuit which by use of stacked crystals and retuning of plate circuit can be made to generate currents of a frequency which corresponds to that of any crystal employed in the stack.

SUMMARY—A discussion of piezo-electric crystals and the early history of the development of the art has been given.

The development of crystal-controlled vacuum tube oscillators by the Naval Research Laboratory has been outlined and various means of amplifying the output of a crystal-controlled oscillator are cited and the best method is described. This method consists of balancing or neutralizing the various stages of amplification and also observing proper precautions for reducing grid circuit losses by using high values of biasing voltage.

A complete high-power low-frequency crystal controlled transmitter is described and a schematic wiring diagram of circuits employing in this transmitter is shown. Schematic wiring diagram and illustrations of one type of low-power high-frequency transmitter complete the subject matter covered in this paper.

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[11]Taylor U. S. Patent No. 1,584,490—May, 1926.

[12]Taylor U. S. Patent No. 1,578,296—March, 1926.

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SIMULTANEOUS PRODUCTION OF A FUNDAMENTAL AND A HARMONIC IN A TUBE GENERATOR*

By
Hoy J. Walls

Assistant Electrical Engineer Bureau of Standards

Methods are available for the transmission and reception of two or more frequencies from a single antenna. [1] The published methods, however, contemplate independent modulation of the several frequencies and require a separate generating tube for each frequency. The method here described involves only a single tube. The application immediately in view was the simultaneous transmission of several standard frequencies; other applications are pointed out below. The work, which was done under the direction of Dr. J. H. Dellinger, was part of the standard frequency transmission program of the Bureau of Standards. The reader should understand clearly that the method is one of multiplex frequency transmission but not of multiplex signal transmission, since there is only a single modulation.

The experiments were made in June, 1924, to determine if it were practicable to operate a radio transmitting set on two or more arbitrarily chosen frequencies simultaneously.[1] The results obtained when operating on two entirely independent frequencies were not as satisfactory as desired, but very good results were obtained when operating on two frequencies, one of which was a harmonic of the other.

The circuit arrangement used is given in Fig. 1. It is similar to the usual "Hartley" circuit but has an additional tuned circuit (L_2C_2) in series with the main tuned circuit (L_1C_1). The antenna circuit is similarly arranged (L_3C_3). It was found that, when the tube was generating a fre-

* Published by permission of Director, Bureau of Standards.

[1] See "Multiplex Radio Telegraphy and Telephony," by Ryan, Tolmie, and Bach. I. R. E., 8, p. 451; 1920.

* Received by the Editor July 19, 1926.

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quency approximately $f_1 \frac{1}{2\pi} \sqrt{L_1 C_1}$ and the circuit $L_2 C_2$ were tuned to some harmonic of f_1 that harmonic hf_1 would be materially amplified. The strength of the harmonic could be readily controlled by varying the coupling between $L_1 C_1$ and $L_2 C_2$, this coupling being the portion of L_2 common to both circuits. If the antenna, whose reactance has been

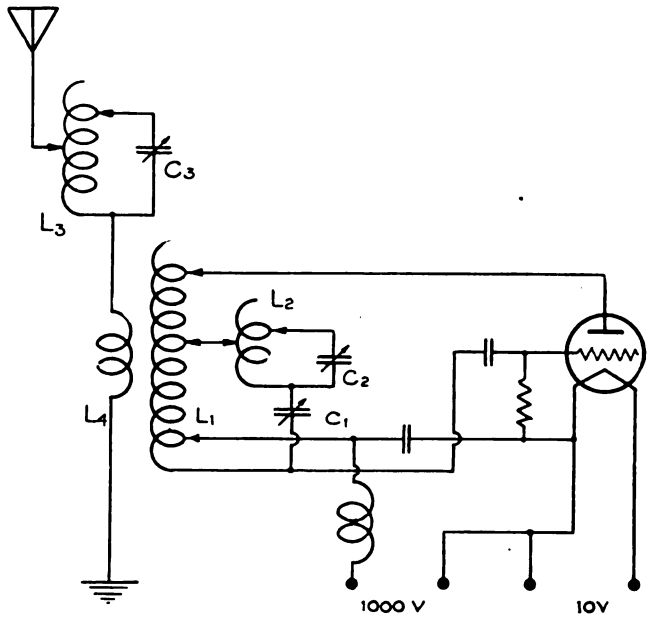

Figure 1—Arrangement for Producing Fundamental and One Harmonic.

adjusted [1] so that it is zero for the two frequencies f_1 and hf_1, is coupled to the inductance L_1 both frequencies are radiated.

Such a tube arrangement, operating on a fundamental frequency of 600 kilocycles with the third harmonic (1800 kc.) strengthened, was coupled to an antenna tuned to 600 and 1800 kc. Good transmission of both frequencies was obtained over short distances. The transmission was not tried over long distances, since a low power tube was used while making the experiments.

Harmonics as high as the tenth (500 kc.) were strengthened with this arrangement when working with a fundamental of 500 kc. Higher harmonics could not be obtained in this case as the limit of the second oscillatory circuit used

[1] See Bureau of Standards Circular No. 74, pp. 41 to 68.

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was slightly over 5000 kc. With a fundamental 2500 kc., harmonics as high as the fifth (12,500 kc.) have been amplified, which was the highest frequency to which the oscillatory circuit used could be tuned. Higher harmonics could probably have been obtained if the oscillatory circuit could have been tuned to higher frequencies. It is to be understood that only one harmonic was amplified at a time. There was no apparent increase in the strength of the other harmonics.

It is evident that the operation of this circuit arrangement depends upon the feedback principle. By inserting the second *LC* circuit in series with the main tuned circuit and tuning the second *LC* circuit to a harmonic, that harmonic would be strengthened somewhat by virtue of the tuning. The then slightly increased harmonic voltage is fed back on the grid of the tube and is amplified, the strength to which the harmonic is amplified being dependent upon the coupling between the main circuit and the second *LC* circuit.

There are several important applications of this circuit arrangement. In relay broadcasting, the main station transmits the same program on two frequencies by means of two independent sets. One frequency is the regular operating frequency in the broadcasting band and the other a much higher frequency, usually between 2500 and 7500 kilocycles. With the circuit arrangement described here it would be quite possible to transmit two frequencies on the same set with any desired output within the capacity of the set on either frequency. This would result in considerable reduction in operating expenses as well as making an additional transmitting set unnecessary.

In calibrating a wavemeter by the use of generator harmonics a generating set of this type is useful when the wavemeter indicator is not sufficiently sensitive to respond to the weak harmonics of the usual generating set. With a fundamental frequency within the range of the wavemeter, any harmonic or higher frequency can be simultaneously obtained with a strength sufficient to operate the wavemeter directly. By changing the fundamental frequency somewhat a new harmonic will be obtained and an indefinite number of points thus secured. To obtain lower frequencies the generating set is adjusted so that the harmonic amplified is within the range of the wavemeter and the fundamental will give the lower frequency.

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DISCUSSION ON REDUCTION OF INTERFERENCE IN BROADCAST RECEPTION (GOLDSMITH)

J. C. Van Horn: The curve of percentages of complaints shown in Figure 10 of the paper under discussion, with its peaks and valleys, seems to me to correspond to what might be called the periods of diminished and normal reception conditions. It is therefore related to the electrical field strength curves of distant stations such as have been described by Dr. Heising in his papers on the measurement of signal strength. I feel sure that if field strength measurements had been made on a distant station (say 1,000 miles from WJZ at Bound Brook) it would have been found that when the field strength was least from distant stations in territory normally served by WJZ, the percentage of complaints was highest. Conversely, when the field strength of distant stations was high, few complaints should have been received.

During the period covered by the actual complaints, the "radio weather" in the Eastern Pennsylvania territory was, I am afraid, very poor. The field strengths from distant stations were so low that radio dealers in the territory in question were at their wits' ends to keep both new customers and existing listeners in a contented frame of mind. The customers complained almost universally that, "the only station that we can hear is WJZ". However, in reality had it not been for the existence of WJZ, very few would have heard any broadcasting at all under such adverse conditions.

*Received by the Editor October 25, 1926.

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AN AUTOMATIC FADING RECORDER*

By

THEODORE A. SMITH AND GEORGE RODWIN

(Technical and Test Department, Radio Corporation of America.)

A device for automatically recording signal intensities is a most useful piece of modern radio laboratory equipment. The value of this apparatus is at once apparent when long, continuous records of signal strength are desired for fading study or for tests of a similar character. It is almost impossible for one man to take a manual record for periods exceeding one hour, when using the ordinary Shaw instrument. Even for shorter intervals the work of keeping a pointer centered on a meter needle is arduous, imposing considerable strain upon the operator. An automatic recorder may be set and left alone to operate indefinitely. Automatic records are also in general more reliable than manual records as the human element is entirely eliminated.

The power required to operate the galvanometer of a Shaw recorder is 0.45 microwatts. Obviously this limited power is not sufficient to operate any ordinary type of graphic galvanometer except one in which the recording is made by a photographic process. Such an instrument is advantageous in that it may be made to follow very rapid fading, but it is inconvenient to use because the fading cannot be observed visually, as well as expensive to construct and operate.

Some means was therefore sought to increase the power available from an ordinary receiver sufficiently to operate a recording meter with a reasonably wide scale, so that fading variations would be well spread out. The use of a radio frequency amplifier of several stages was considered, but was not attempted because of the difficulties in getting the number of stages necessary, to operate stably.

It was desirable that the fading curves be proportional to the carrier field intensity. If a constant note modula-

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tion could be provided at the transmitter, the resulting audio frequency output of a receiver could be amplified sufficiently to operate a recording milliammeter. However, when fading records are desired, of radio broadcasting stations, constant tone modulation is seldom available, and even if such was the case, the resulting curves would show fading of both carrier and sidebands together. An alternative is to supply the constant note modulation at the receiver and to filter out any other modulation which may have been impressed on the carrier at the transmitting station. An audio frequency power amplifier could then furnish enough energy to operate a relatively insensitive graphic meter.

Such an arrangement was suggested by Mr. Julius Weinberger of this Department, and it has been employed with entirely successful results. Two sets were constructed, one for use on broadcasting frequencies only, the other for covering from 16.6 megacycles to 2 megacycles (18-150 meters) by the use of plug-in coils. With an external oscillator tube, the range of the latter was extended to the band from 300 to 100 kilocycles (1000-3000 meters).

The two sets are similar electrically, each consisting of—a superheterodyne receiver, a 1000-cycle modulator, an audio frequency power amplifier, a 1000-cycle pass filter, a rectifier, and a recording milliammeter. The broadcast receiver is a standard Radiola 28, eight-tube set. The short wave outfit consisting of a Standard six-tube "catacomb" such as is found in the Radiola 25, but with the high-frequency amplifier and oscillator external. A General Radio 1000-cycle hummer furnishes the modulation which is introduced into the grid circuit of the radio frequency tube, as shown in Figure 1. The intermediate frequency amplifier, second detector and two stages of audio frequency amplification are contained in the "catacombs" of the receivers. A parallel circuit tuned to 1000 cycles is placed between the second detector and the first audio amplifier. This has been found quite satisfactory in removing the broadcast modulation to a degree insufficient to operate the recorder. A power stage of audio amplification feeds into a rectifier to which the meter is connected.

The recording milliammeter was furnished by the Esterline-Angus Company and is of the electrodynamometer type. The power required to operate it at full scale with all coils

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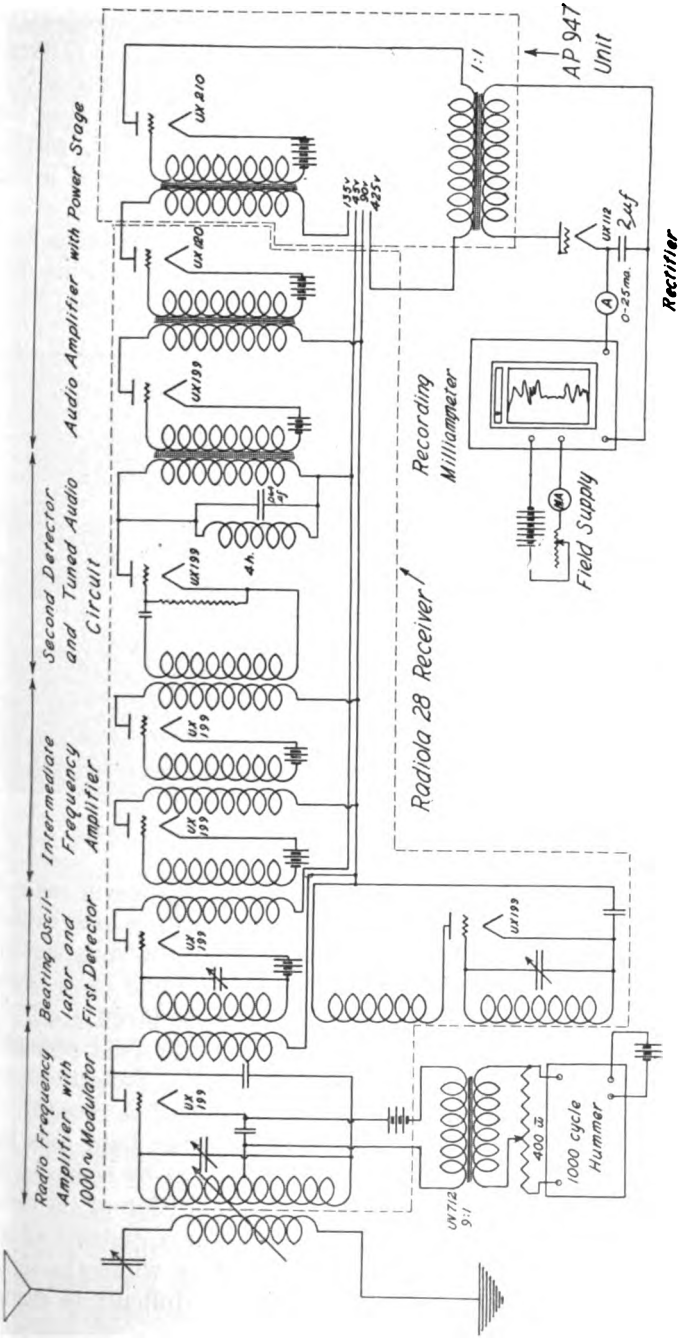

Figure 1

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in series is about three watts. Its resistance under the same conditions is 32, 940 ohms. Full scale current is 10 milliamperes. The meter can be used at record speeds varying from 60 feet (18.3 meters) per hour to ¾ inch (1.9 centimeters) per hour, and thus either short or long-period fading records can be obtained. A second pen marks along the margin of the record and by closing a local circuit, indentation may be made on this trace for the purpose of noting time or disturbances on the record.

With all coils in series, the fading record ordinates were roughly proportional to the fourth power of the field strength owing to the square law action of the detector and

Figure 2—Two Automatic Fading Recorder Sets

the use of a dynamometer type meter. However, as now used, the meter field current is supplied from a local battery and thus the ordinates are proportional to the square of the field strength except at the extreme lower end of the scale. As the record paper is printed with a squared scale, the printed ordinates are proportional to the field strength. This is a convenient arrangement, as the deflections are roughly proportional to the audio output of a receiver.

The use of direct current in the meter field coils has proved advantageous for changing scales, as by reducing the field current by a known per cent, the deflection of the meter can be reduced in the same ratio. This is a distinct advantage where signals are to be compared or when the signal is fading considerably and it would be difficult to change

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scale in a known proportion by any other means. With one set of coils supplied from a direct-current source, the power required to produce full scale deflection on the meter is reduced to one-half, or about 1.5 watts, which may readily be obtained from a UX210 power tube operated properly.

A six-volt storage battery furnishes the filament supply for all tubes except the power amplifier which is run from

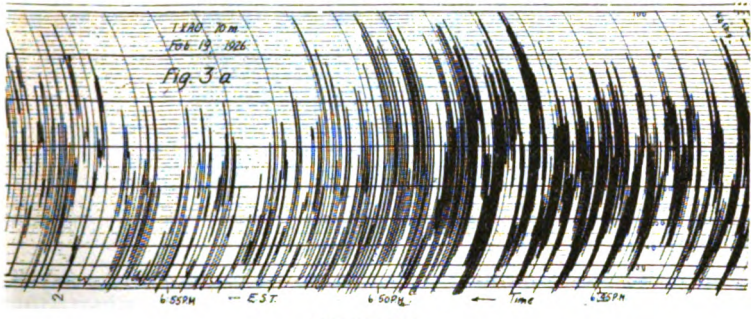

Figure 3a

alternating current. The B battery current is supplied by the power amplifier unit, shown under the tables in Figure 2.

Provision is made for using a manual recorder in the sets constructed. For measuring field intensities, known

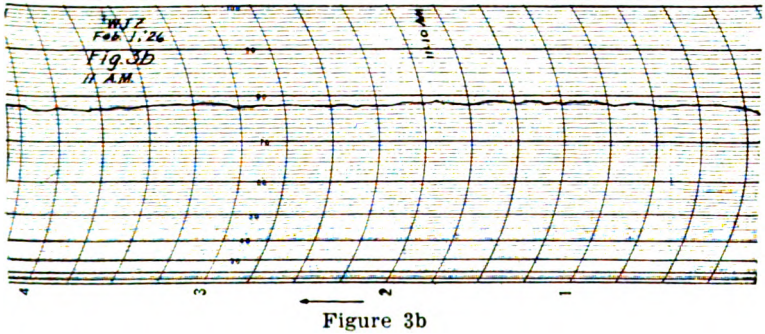

Figure 3b

radio frequency voltages may be introduced into the input circuit.

Sample records taken at various times are shown in Figures 3 and 4. All reception was at the Radio Corporation Technical Building at Van Cortlandt Park, New York City. Figure 3a is a portion of a continuous record taken

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in the evening of February 19, 1926 of station 1XAO located at Belfast Maine, transmitting on 70 meters. (Belfast is located 350 miles or 560 kilometers from New York City). This record shows the sudden change in period fading some-

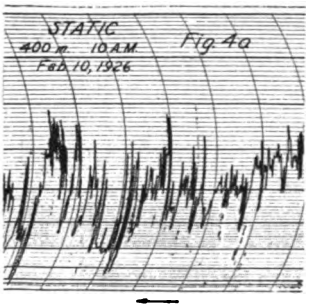

Figure 4a

times encountered on short waves. Figure 3b is a daylight record of WJZ at Bound Brook, New Jersey, transmitting on 454.3 meters (660 kilocycles).

Figure 4a is a record of static on a wavelength of 400

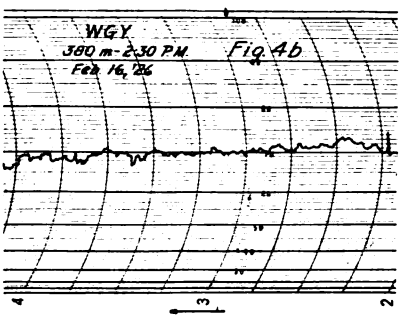

Figure 4b

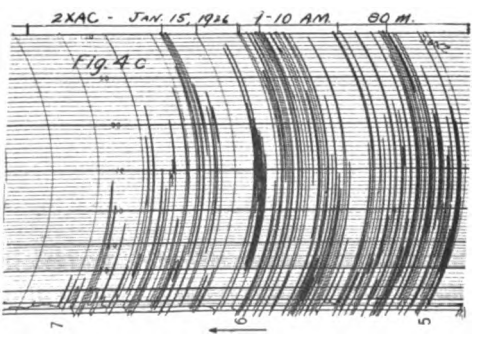

Figure 4c

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meters (750 kilocycles) during a snow storm on February 10, 1926 at 10 A. M. Figure 4b shows fading encountered in New York City from WGY in Schenectady in daylight

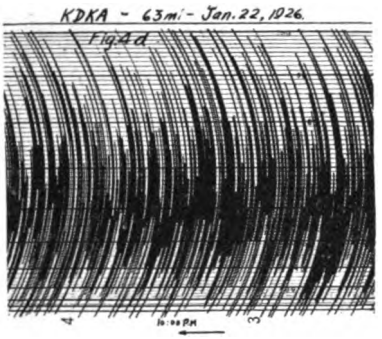

Figure 4d

on 380 meters (790 kilocycles). The record in 4c is of short wave transmission on 80 meters (3750 kilocycles) from Schenectady, also in daylight. Figure 4d is transmission from Pittsburgh on 63 meters (4760 kilocycles) at night.

SUMMARY—A device for automatically recording signal intensities is described with the method employed to amplify the signal sufficiently to operate a commercial type of graphic meter. Sample fading records of various transmissions are presented.

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BEHAVIOR OF ALKALI VAPOR DETECTOR TUBES*

By
Hugh A. Brown and Chas. T. Knipp
(University of Illinois)

I—Introduction

Since the original investigation of certain alkali vapor tubes used as detectors† was completed, new and more sensitive types of tubes have been developed and put on the market. Interest has centered around the comparative efficacy of the later tubes with the supersensitive (potassium sodium alloy) tubes previously developed. It is the intention of this paper to point out not only the actual comparisons, but also the manner in which the comparative tests were made. In the past, erroneous interpretation of comparative results have often occurred, due to variations in the manner of conducting tests.

Some very peculiar and interesting features of behavior of the alkali tubes will also be described and illustrated. (These unique features will probably be the ultimate basis for a scientific explanation of the physical phenomena occurring within tubes of this type.) The observations are confined to tungsten filament tubes into which the molecular alloy of potassium and sodium has been distilled at a temperature of 250 deg. cent. by the oil bath described in Bulletin 138‡. This particular treatment seems to yield finer results in detection than any other single alkali metal vapor content. A few observations were also made of the effect of forcing sodium through the glass walls, this latter treatment yielding no sensitive detection performance whatever.

II—Transient Plate Current Variations

In order to produce a sensitive potassium sodium alloy content detector it is essential to use a tungsten filament

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†See Bulletin 147, Eng. Exp. tSa., University of Illinois.
‡See Bulletin No. 138, Univ. of Ill., Eng. Exp. Sta., also Proc. I. R. E. Vol. 10, No. 6, p. 451.

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carrying 0.7 to 1.0 ampere which will produce sufficient heat to make the tube walls fairly warm, say 40 deg. cent. on the outside. The tube is not stable in its action until the tube envelope gets hot, and this occurs in about one minute after the filament current is turned on. As soon as the tube begins to warm up the plate current rises, slowly at first, increasing its rate of increase as the tube gets warm; then ceases to rise, and finally decreases to a steady value when the tube walls have warmed up to the final temperature. This phenomenon is shown in Fig. 1. The response of the

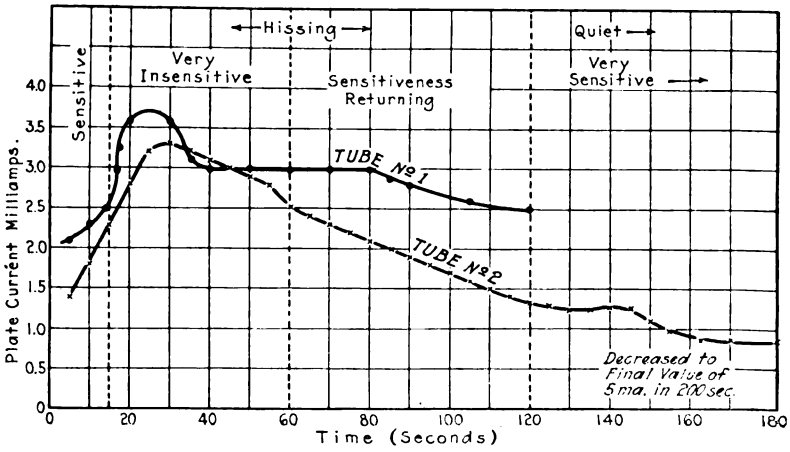

Figure 1

tube as a detector is rather peculiar and is also indicated on the curve sheet. The peculiar results shown in Fig. 1 do not obtain for any type of high-vacuum tube or gas-content tube now or formerly in general use. Whether or not the tube be well or poorly outgassed when the alloy is introduced, the results are the same in a general way. If there is considerable residual gas the variation is greater as the tube warms up than when the tube is well outgassed, however the writers have not been able to produce tubes which do not show this peculiar quality, and are therefore not able to say whether or not it is due to the presence of residual gas. While the plate current is increasing, a hissing sound occurs in telephone receivers placed in the plate circuit of the tube, the sound increasing in intensity to a maximum, then gradually reducing again until it is not heard when

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the plate current has reached its steady value. The more "gassy" the tubes the more intense is the hissing sound.

The cause of the variation in plate current detector performance has not been found. It is certain that as these quantities vary the vapor pressure of the alloy within the tubes is increasing until the tube walls reach a final temperature. The increasing vapor pressure in some way affects the behavior noted. The exact mechanism of this, is however not understood at present.

III—Performance Characteristics

1. *Variation of Response with Plate Voltage, and Input Voltage.* When a high vacuum tube is used as a detector the optimum response will occur at low plate or anode bat-

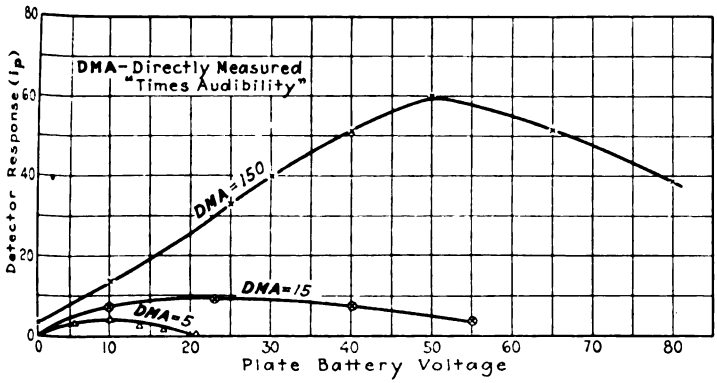

Figure 2

tery potential when the input voltage is very small. When the input voltage is comparatively very large, producing an optimum response having a directly measured audibility of about 150 times, this optimum response occurs at the highest plate voltage that can be provided just below the point where residual gas atoms become ionized. This is usually about 80 to 100 volts for the small amplifier tube now in use. A soft detector tube will not respond at medium plate voltages at all, due to intense ionization. But potassium sodium alloy tubes respond to strong input or signal voltages with increasing intensity as the plate voltage is raised as do high vacuum tubes, even though these tubes contain considerable "inert" gas, as a result of incomplete evacuation. Their response under such conditions is greater than for all other

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commercial tubes tested except for the UX—200—A. This will be discussed later.

The behavior of a potassium sodium alloy detector under

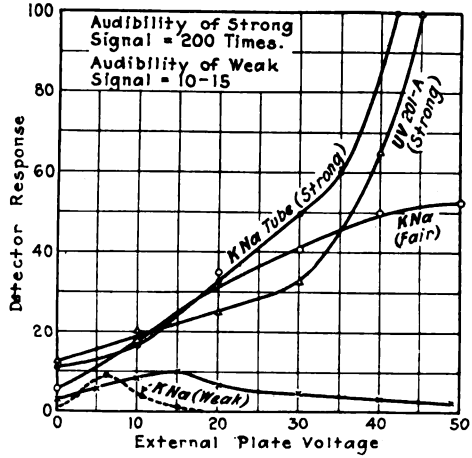

Figure 3

conditions of strong and weak signal voltages is shown in Fig. 2. This property is also possessed by the conventional high-vacuum detector-amplifier tube. The condition of low

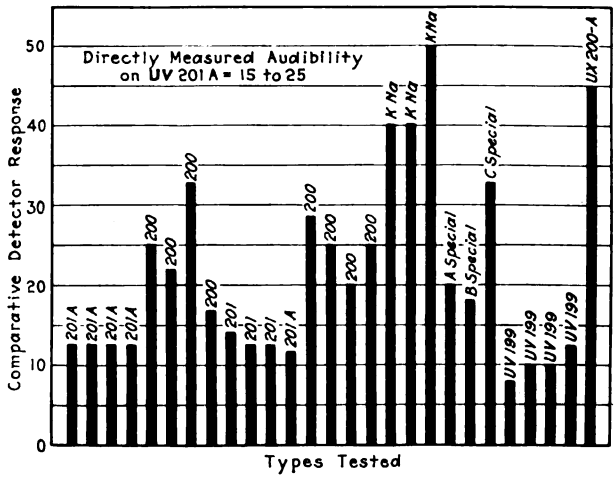

Figure 4

optimum plate voltage is obtained for a much weaker signal voltage in the case of a potassium sodium alloy tube than for the conventional high-vacuum type. This is illustrated in

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Fig. 3. Not only is the potassium sodium alloy tube a more efficient detector at low plate potentials and low signal voltages, but it is also a more efficient one at high plate potentials (40 to 50 volts) and strong signal voltages. The comparison is made using the optimum plate voltages for the conditions of a strong and weak signal voltages, according to the curves of Figs. 2 and 3.

2. *Comparative Efficiencies.* Fig 4. shows the results of comparative tests on various typical tubes to illustrate the present comparative efficiency of the potassium sodium alloy detector. Since making this test the Radiotron UX—200—A has been put on the market and one typical com-

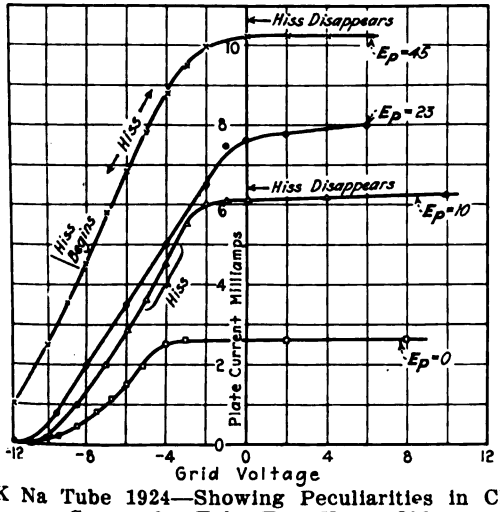

Figure 5—K Na Tube 1924—Showing Peculiarities in Characteristic Curves for Tube Two Years Old

parison is added, showing it to be about as efficient as the potassium sodium type, but requiring optimum plate voltages of 40 to 50 as compared with 10 to 20 volts for the potassium sodium tubes on weak signal voltages. For strong signal voltages also higher plate voltages are required for the same degree of response as the potassium sodium tubes give.

3. *"Hard" and "Soft" Tube Peculiarities.* Attention is again called to the fact that increasing the filament temperature gives the "gassy" potassium sodium alloy tube characteristics of a high vacuum amplifier (See Bulletin 138 loc. cit.). This is especially true of the shape of the characteristic curves of such tubes also illustrated in the

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bulletin referred to. As a matter of interest Fig. 5 is given showing how a potassium sodium alloy tube behaves during the change in grid potential indicated by the abscissa of the curves. While the data taking operation is proceeding a hissing sound occurs in phones in the plate circuit, then the ranges of E_g indicated, as if ionization were setting in, but the curves are not "kinky" as those of gas content tubes, and for higher values toward $(+E_g)$ the hissing ceases. This tube in use gives both the sensitive detector performance on weak and strong signal voltages, and the efficient amplification of strong amplifier input voltages. In either service it is a considerable improvement over the same tube not

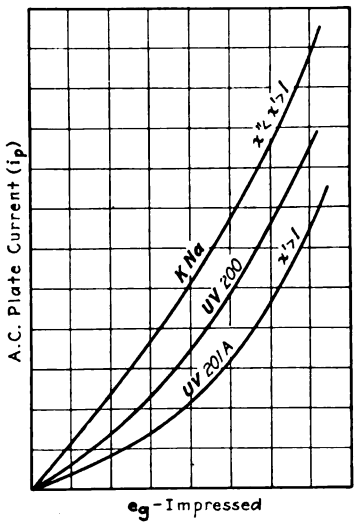

Figure 6

primed with the alloy. The degree of rise and fall of plate current during the warming up process and the attendant hissing noise were decreased by more thoroughly outgassing the electrodes with the aid of a high frequency induction furnace. However, such tubes were seemingly just as sensitive as those which were more "gassy." The transient plate current characteristic shown in Fig. 1 could not be eliminated by heat treatment and long continued evacuation, hence it seems that the transient "warming up" current and hissing is due to increasing activity of the alloy molecules as the alloy vapor pressure increases.

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IV—"Square Law" Variation

It has been generally assumed that thermionic tube detectors obey a "square law" relation, that is, the a-c. component of plate current i_p is: $i_p = K\ (e_g)^2$ where e_g is the impressed or signal voltage on the grid circuit. In general $i_p = K(e_g)_x$ and for potassium sodium alloy tubes repeated variations showed that x is more nearly unity than for either "soft" or "hard" detectors. This is illustrated in Fig. 6. This should mean quite an improvement in reception of radiotelephone currents, as it would give more nearly the desired condition of the ideal linear rectifier.*

V—Conclusion

Further work has shown that alkali vapor detector tubes, especially those containing the molecular alloy of potassium and sodium, are ideal tubes for durability, true tone reproduction and non-critical adjustment of plate and filament voltages.

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DISCUSSION OF A METHOD FOR MAXIMIZATION IN CIRCUIT CALCULATIONS (ROBERTS)

O. C. Roos: Mr. Roberts' interesting paper illustrates a well known theorem concerning the "Hodograph" of Sir Wm. Rowan Hamilton—the discoverer of quarternions.

The Hodograph is simply a polar curve of velocities, which is capable of being divorced from all "imaginaries" for the sake of graphical or mechanical analysis, using the parallelogram of velocities.

"Maximization" of velocities and hence of all other functions changing with time can by its use, be rendered perhaps physically clearer than is possible by the use of imaginaries, as I shall attempt to explain and illustrate below. But first a word on the Hodograph itself is necessary.

In Fig. 1*a* is shown the path of a particle which travels from point *a* to point *g* along the curve *a b c d e f g*. Its

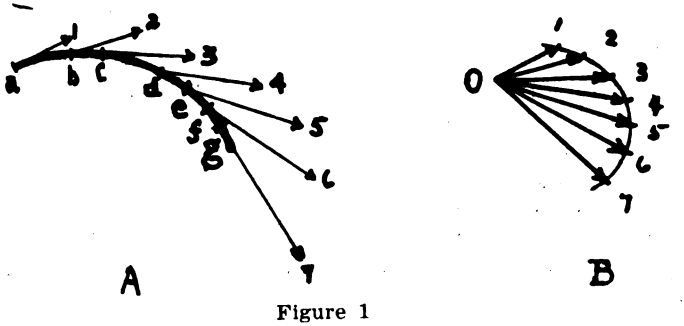

Figure 1

velocity at each of these points is given in magnitude and direction by the arrows *a*-1, *b*-2, *c*-3, *d*-4, *e*-5, *f*-6, *g*-7.

If we now collect all these arrows as vectors with their initial points at *O* in Fig. 1*B*, keeping their directions unchanged, we have the Hodograph 1-2-3-4-5-6-7, whose sectors 1-2, 2-3, etc. 6-7 are a measure of the change of velocities of the particle as it passes between the corresponding points *a*-*b*, *b*-*c*, etc. *f*-*g*, in Fig. 1*a*.

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In the limit, the Hodograph is a smooth curve and the velocity at the end of the arrow (vector) from O as it passes from points 1 to 7 is an instantaneous measure of the direction and magnitude of the velocity change, i. e. the acceleration.

Now the applicability of the Hodograph to Mr. Roberts' problem has long been known and is analyzed and applied as follows. The fact is that any monogenic function can be

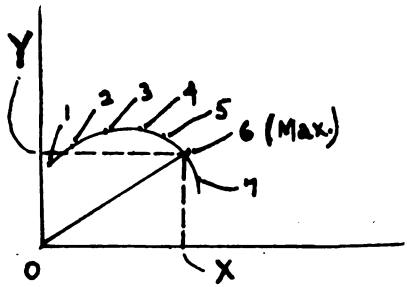

Figure 2—Analysis of Hodograph into Perpendicular Components.

continuously resolved into two components at right angles to each other and its growth in time is proportional to their rate of exchange in time.

If we turn Fig. 1*b* around and redraw it in Fig. 3, as a polar curve 1-2-3-4-5-6-7, which is z, a function of ox (real) and oy (imaginary) we may consider both ox and oy as dependent variables, both being functions of time. Hence we may consider them as velocities.

Let us postulate that the velocity at 06 is a maximum. Hence the next elementary change in velocity 6-7 must be right angles to 0-6 and measures the acceleration of this maximum velocity. Then in Fig. 3 we decompose the velocity 06 into an X-component OX corresponding to energy cyclically consumed, and into a Y-component OY, corresponding to energy cyclically stored.

We now can lay off a shorter "maximum" velocity OX along 06, making OX' equal to OX. It is now obvious that the relative acceleration of OX' is the same as 06, since the relative accelerations, 6-7 and X'—8' are the same. Hence the X-component, if it has been changing with time, must not change toward or away from R, but the acceleration must be at right angles to OR, from X, i.e. it must equal X—9 which equals X—8.

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Thus the X component is "stationary" or its rate of change or acceleration is zero when X is the variable. A similar argument shows that the Y-component must be "stationary" when Y is the variable.

Of course the device of considering a physical quantity like an impedance, as a velocity, has *a priori* no real justifi-

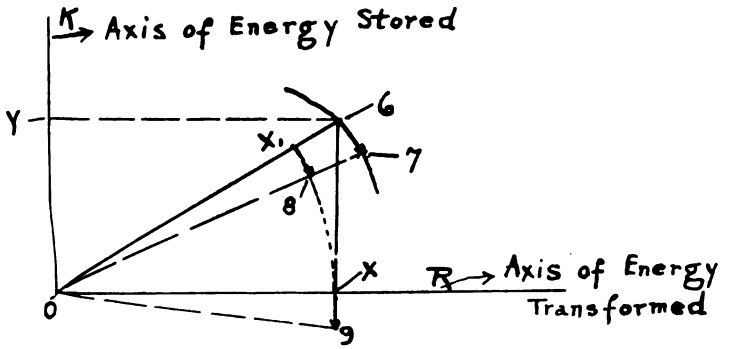

Figure 3—Decomposition of Physical Acceleration into Wattful and Wattless Components.

cation. It is just a convenience. However, we have the coincidence that impedances in the A. E. M. system are velocities, an ohm being 10^9 cm. per sec. whether wattful or wattless.

For those who are not familiar with vectors the Hodograph is thus an aid to physical visualization of a polar maximum.

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REPORT CONCERNING THE OBSERVATION OF THE INFLUENCE ON THE PROPAGATION OF RADIO-WAVES OF THE SUN ECLIPSE OF THE 14th OF JANUARY 1925 IN THE DUTCH EAST INDIES

By
E. C. HOLTZAPPEL

In order to observe the influence on the propagation of radio-waves of the sun eclipse of the 14th of January this year, which lasted at Weltevreden (Batavia) from 0603 G. M. T. till 0844 G. M. T., the wireless stations at Malabar (PKX on 15.6 km., ANA on 7.7= km. wave) and Tjililin (ANF on 30=Meter wave) both near Bandoeng (Java)—PKX and ANA at about 26 km. south, and ANF at about 18 km. west from Bandoeng—were ordered to transmit simultaneously on the 13th, 14th and 15th signals on 15.600, 7700 and 30 meters during the period 0300 G. M. T. till 10.00 G. M. T.

The first wave was used by the 2400=kw. Poulsen arc (PKX), the second by the high frequency alternator of 800=kw. (ANA) and the third by a 5=kw. short-wave transmitter at Tjililin (ANF).

The foreign stations at Tananarive (Madagascar), Guam (Pacific), Cavite (Philippines) and Saigon (French Indo-China) were asked to listen in to those signals and to report the results and if possible transmit themselves during the time of the eclipse for observation at Bandoeng.

Further orders were given to listen in to the stations in these Colonies at Sabang, Padang, Bengkalis, Palembang and Sitoebondo, whereas our station of Bengkalis was asked to transmit on her Poulsen arc of 6 kw. on 1650 meter wave; furthermore the private stations at Tarakan and on board ships were requested to observe the Bandoeng signals. Signals of different stations were observed at our receiving posts at Rantja-ekek and Padalarang, both near Bandoeng and belonging to the Malabar system.

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General results of these observations have been negative as was anticipated, that is, the observed influence of the eclipse on the propagation of radio signals has been almost inperceptible or did not exist at all.

For the stations at a great distance from the track of the eclipse shadow on earth this could be expected, but even at Palembang and Tarakan, lying directly in this track, nothing has been observed of any such influence.

The only exceptions were observed at Rantja-ekek with regard to the signals of Tananarive on 15600 m. and of Bengkalis on 1650 m. which were actually weaker during the eclipse.

On the contrary the signals of Saigon at Rantja-ekek, and the Malabar signals at Sitoebondo during that period were much stronger than ordinarilly which indicates reflection from the shadow zone.

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INSTITUTE ACTIVITIES

Annual Convention

For three days beginning January 10th, the Annual Convention of the Institute was held. The attendance at all of the sessions was very good. At the opening meeting on the 10th the new officers for 1927 were installed and the Liebmann Memoriay Prize for the year 1926 was presented to Dr. Ralph Bown. There followed a paper on "The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism" presented by Dr. G. W. Pickard. Much discussion followed. In the afternoon of this day many of the convention attendants visited the American Telephone and Telegraph Company's station on Walker Street. At 8 P. M. Dr. E. F. W. Alexanderson gave a talk, illustrated with apparatus set-ups, lantern slides, and moving-pictures on the General Electric Company's progress in transmission of photographs at a high rate of speed. The auditorium was packed to capacity at this meeting. On January 11th a paper by Professors Hugh A. Brown and Charles T. Knipp on "The Behavior of Alkali Vapor in Detector Tubes" was read. A series of informal discussions followed. In the afternoon trips were made to the Electrical Testing Laboratories, Dubilier Radio & Condenser Corporation, and the Federal Brandes Products Corporation. At 8 P. M. on this evening a paper by D. K. Martin on "Simultaneous Transmission of Broadcasts by Chain Stations on the Same Wave Lengths" was read by Mr. Martin. The last day of the convention, January 12, started off with four inspection trips; the Radio Corporation of America's transoceanic terminal apparatus station, the Driver-Harris Company, studio and control room of WEAF and the studio and control room of WJZ. In the afternoon a paper by W. A. McDonald was read. The subject of this paper was "The Importance of Laboratory Measurements in The Designs of Radio Receivers." The convention closed with the banquet at the Waldorf-Astoria where about three hundred members were present. Mr. D. Rigney acted as toast-master introducing one form of light entertainment after another until the session was officially brought to a close at midnight.

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Luncheon for Retiring President

On the first day of the Convention, January 10, a luncheon was given at the Fraternity Club to retiring president Donald McNicol. The affair was attended by six past-presidents and by officers who were in attendance at the first day's sessions. Those present were: Donald McNicol, Dr. M. I. Pupin, Dr. J. H. Dellinger, Dr. Ralph Bown, Dr. G. W. Pickard, J. V. L. Hogan, Lloyd Espenschied, Laurens E. Whittemore, Raymond A. Heising and R. H. Marriott.

January Meeting of Board of Direction, January 5, 1927

At the meeting of the Board of Direction, held at Institute Headquarters on January 5th, the following were present: Donald McNicol, President; Dr. Ralph Bown, Vice-President; R. H. Marriott, J. V. L. Hogan, W. F. Hubley, and J. M. Clayton, Assistant Secretary. The members of the Board counted the ballots in the annual election of officers, the result being as follows: for President, Dr. Ralph Bown; for Vice-President, Frank Conrad; for Managers, R. A. Heising, and J. F. Dillon. Three additional managers are to be appointed at the February meeting of the Board to fill the terms of the three members leaving the Board.

The following were transferred to Member grade: Wilson Aull, and Clair D. Mitchell. The following were elected directly to the grade of Member: A. G. Lee, J. R. Land, Valerian Bashenoff, J. L. Bain, Jr. R. Martin. L. G. Pacent was transferred to the grade of Fellow.

One hundred and fifty-eight Associate members and 19 Junior members were elected. Treasurer Hubley submitted a statement of the disbursements for the year 1926.

Sections Committee

The Sections Committee, D. H. Gage, Chairman, held a meeting at the Institute Headquarters on January 11th. A number of out-of-town Section Chairmen and representatives of different Sections were present. Many matters of importance in the organizing and managing of Sections were discussed. Much work was done toward standardizing the procedure to be followed in establishing a Section.

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News of the Sections

Los Angeles Section

The Los Angeles Section held a meeting in the Commercial Club of Los Angeles on December 20th. Mr. Edward H. Gilford presented a talk on "Locating Unknown Metal Deposits by Radio, or Re-Radiation Applied to the Locating of Unknown Conductors." A general discussion followed. Forty-seven members attended the meeting which was preceded by a dinner.

Canadian Section

The Canadian Section held a meeting on the evening of January 5th at the Electrical Building, University of Toronto. Mr. Hepburn, Chairman, presided. Mr. V. G. Smith of the University of Toronto, presented a paper on "A Mathematical Study of Radio Frequency Amplification." This paper was discussed by C. I. Soucy, A. M. Patience and others. Due to the resignation of G. F. Eaton, Mr. Patience was elected Vice-Chairman. There were forty-six members present.

The next meeting of the Section will be on February 2nd in the Electrical Building of the University of Toronto at which time Mr. R. H. Langley will present a paper "Investigation Into Broadcasting Conditions."

Appointment of Assistant Secretary

The Board of Direction of the Institute has employed Mr. John M. Clayton as Assistant Secretary of the Institute. Mr. Clayton will be permanently located at headquarters, 37 West 39th Street, New York, and will have direct charge of office management, records, correspondence, and all matters pertaining to Institute activities.

Mr. Clayton has for some years been assistant technical editor of QST Magazine, of the American Radio Relay League, Hartford, Conn. He is well versed in radio engineering literature, is familiar with the business of radio organizations and has a wide acquaintance among radio engineers throughout the United States and Canada. He took up his duties with the Institute beginning January 1, 1927.

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OPENING ADDRESS OF DONALD McNICOL RETIRING PRESIDENT

Annual Meeting and Convention, Institute of Radio Engineers, January 10, 1927.

This meeting is the opening session of the fifteenth annual meeting of the Institute of Radio Engineers. Beginning in 1926 the annual meeting was conducted in connection with a Convention continuing over several days. The 1926 Convention was the first gathering of the kind carried out by the Institute. The meeting was an outstanding success and it is likely that the Conventions will become permanent annual events.

From one standpoint it can be imagined that the public may view with little patience the holding of a three-days convention by radio engineers. There is no other art, no other science, in which the public has taken or is taking so keen an interest, such a proprietary interest, as in radio. Even the Doctors of Medicine, and surgeons, who deal with the affairs of physical well-being—with life itself—are permitted to carry on their researches in their own constructive, if leisurely way, without being assailed as delinquents. And, when the best these savants can do to outwit parasitic influences, and to destroy malignant interferences, fails, they are forthwith absolved on the ground that the purposes of destiny or fate must be served.

The public demands of radio engineers solution of electrical and meteorological problems involved in establishing and maintaining year 'round, satisfactory radiophone transmission and reception.

Had the popular use of radio for entertainment purposes not been precipitated at a time when but a small number of engineers had become proficient in the art it is probable that advance for a time would have been conducted along lines other than that of broadcasting.

It is true, of course, that the great pressure brought upon the engineer to unravel the difficulties of electrical transmission over an uncontrollable medium have served to hasten desirable results. And, measured by the achievements of the past five years much has been accomplished. To realize the extent of the advance it is necessary only to

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view the best of the equipment in use five years ago and to recall the cacophonous emenations vibrated by the horns of 1921 into the virgin air.

However, we shall hope that the public will be indulgent and await the close of this convention before demanding More and Better Radio.

Some day one of us may have the leisure to write a story about the intellectual rise in radio. If the job is well done it is likely that a very interesting picture will be developed recording the purely electrical advance up to the point where a metal conductor between the source of power and the device to be operated no longer remained a necessity. Thenceforward the account will be less difficult to prepare historically because of the large amount of written matter available, produced by physicists, engineers and students engaged in carrying on research and experiments.

The elementary but spectacular 'spark' operation for telegraph purposes, for a decade prior to the advent of the Audion, enabled communication engineers to get used to the idea of wireless working.

The chronicler will perhaps need to exercise a sort of journalistic charity while dealing with radio advance from 1906 until 1912, but from the latter year forward when the possibilities of the wonder-working three-eelctrode tube were fully realized, the science has progressed to the stage where every man's house has a pair of ears—or, at least, an ear. It is a magic ear—one not only with which the listener may hear the Sunday sermon a few blocks distant from his morris chair, but with which he purposes to overhear neighborhood gossip taking place in a hamlet across the continent, or across the seas.

Indeed, even now, the secrets of the laboratories leaking out, every man will soon require that he be enabled at will visually to observe what is taking place at any point in the country where his inquiring eye may desire to roam.

A few years ago these developments were of the stuff of which dreams are made. The task of the radio engineer is to make these dreams come true, and the part which the Institute of Radio Engineers has played and shall continue to play in this realization is of the first importance.

In the beginning, fifteen years ago, the Institute was nursed into being by an earnest, enthusiastic, small group of amateurs, who were in fact the engineers of their day.

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From 1912 until 1924, the Institute existed mainly for the purpose of holding periodical technical meetings in New York and for the purpose of financing the publication of radio technical papers; until 1915 groups of papers were published four times each year, and from 1916 until 1926 inclusive the Proceedings have been published six times each year, and with the January, 1927, issue the Proceedings are to be published monthly.

In 1924, President Morecroft introduced and fostered several organization and management betterments which paved the way for the aggressive, constructive administration of Institute affairs carried on by President Dellinger throughout the year 1925. These years witnessed marked gains in Institute development, activities and usefulness.

At the time the Institute was organized in May, 1912, there were less than fifty members. On June 1, 1914, there were 79 Members and 311 Associates. On January 1, 1916, there were 39 Fellows, 108 Members, 764 Associates and 73 Juniors; a total of 984. On January 1, 1922, the total was 1600, all grades, and on January 1, 1926, 2300, all grades. On January 1, 1927, the total membership was approximately 3,800.

The gratifying increase in membership during the past two years has been due largely to further development of the Section idea.

The Washington Section was organized in January, 1914; the Boston Section in November, 1914; the Seattle Section in February 1915, and the San Francisco Section in 1916.

The Chicago and Toronto Sections were organized in the summer of 1925, and in 1926 Sections have been organized in Rochester, N. Y.; Los Angeles, Calif.; Hartford, Conn., and the Philadelphia Section reorganized. During the year 1926, in addition to the eleven New York meetings, forty-nine meetings were conducted by the various Sections, at which technical papers were presented.

PUBLICATIONS

In 1926 a Year Book was published containing a complete list of the membership, together with a considerable amount of other matter of interest to members.

In the Proceedings forty-two technical papers were pub-

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lished, and supplies of pamphlet copies of all papers presented at New York were distributed at the meetings.

Work was begun on an Index of all papers presented before the Institute 1913 to 1926, inclusive.

Information has been gathered for the publication of a Year Book to be printed early in 1927.

Diplomas have been issued to Fellows and Members, and membership cards to Associates.

If among new features of management introduced in 1926, any may be noted as being an innovation, we might refer to the effort put forth to bridge over the gulf which in learned societies often separates the younger men and less experienced members from the advanced engineers and leaders in the science.

There is need and opportunity for further development along this line, but in the past year we have succeeded in having present at New York meetings many of the outstanding leaders in radio science, so that the younger men, largely in the majority, might become acquainted with and listen to oral discussion presented by engineers recognized as authorities.

Both expert and novice have been given opportunity to speak from the floor on the subjects presented at meetings. Already there is evidence that from this good has followed.

After all, the Institute's main purpose is to disseminate engineering information bearing on the present problems of radio, and it is plain that methods followed to set forth and to interpret the laboratory achievements of the physicists are successful and widely useful in proportion to the number of students who leave each meeting with knowledge gained of such advances and with clearer understanding of fundamentals.

Meetings and Papers Committee

The Meetings and Papers Committee during the past year, under the direction of R. H. Marriott, provided excellent papers for all meetings and for the pages of the Proceedings. The committee held regular monthly meetings during the year and carried on a continuous correspondence for the purpose of procuring suitable papers for presentation.

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Membership Committee

The Membership Committee, H. F. Dart, chairman, was thoroughly active throughout the year. Much of the substantial gain in membership is directly due to the work of this committee. Regular monthly meetings were held, usually in the evening, and there were many meetings of subcommittees.

Admissions Committee

This committee, Ralph Bown, chairman, held monthly meetings throughout the year. During the year the committee approved five applications for transfer to Fellow grade; sixty-seven applications for transfer from Associate to Member grade, and fifty-four direct elections to Member grade.

Twenty-four applicants for Member grade were recommended for Associate grade; seventeen applicants for transfer from Associate to Member grade were not approved, and four applications for transfer to Fellow grade were not approved.

Standardization Committee

The Standardization Committee, L. E. Whittemore, chairman, has had a number of meetings during the year, the work done being preliminary to a revision of the 1926 Standards Report, to be published perhaps in 1928.

Sections Committee

The Sections Committee, David H. Gage, chairman, is to be credited with much of the work done toward rehabilitating the activities of the older Sections and organization of the new Sections. The Committee has in hand constructive plans looking to the establishment of one or two additional Sections.

Committee on Publicity

The Committee on Publicity, W. G. H. Finch, chairman, has procured for the Institute a very gratifying amount of space in daily newspapers and radio periodicals. All of the New York meetings of the Institute have had satisfactory

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prior notice in news sections of dailies, and a considerable amount of matter relating to Institute activities has been published in weekly and monthly periodicals. The committee also has had entire charge of the extensive publicity given the conventions, and has attended to details of circularization and printing.

Advertising and News Notes in Proceedings

The agency having charge of the procurement of paid advertisements for the Proceedings has done good work during the year as is evidenced by the number of full-page advertisements appearing regularly in those pages. With the Proceedings on a monthly basis the advertising space occupies should materially increase.

The new feature of incorporating in the first six or seven pages of each issue of the Proceedings notices, reports and news items of interest to members has had a particularly beneficial effect in increasing the interest of Institute members in all parts of the country, and Canada, in the activities of the officers and of the committees.

Presentation of the Liebmann Memorial Prize

The Institute each year makes two awards to radio investigators in recognition of noteworthy inventions or other meritorious developments in radio technique.

One award, the Institute Gold Medal of Honor has been awarded annually since the year 1918. The following engineers have received this prize: E. H. Armstrong, E. F. W. Alexanderson, G. Marconi, R. A. Fessenden, Lee De Forest, John Stone, M. I. Pupin and G. W. Pickard.

The Liebmann Memorial Prize consists of five hundred dollars in cash, being the annual interest on a sum of money presented to the Institute for the purpose of maintaining the award. This prize is awarded by the Board of Direction of the Institute each year to the engineer who has made what is clearly the most outstanding valuable contribution to the art and science of radio during the previous year.

Engineers who have received this prize since it was established in the year 1919, are: Leonard F. Fuller, Roy A. Weagant, Raymond A. Heising, C. S. Franklin, H. H. Beverage, John R. Carson and Frank Conrad.

The elements of mystery bound up in the term "Radio Transmission Phenomena" have a different meaning for

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the average non-technical broadcast listener, than for the radio engineer. The radio engineer is at all times several giant steps ahead of the merely curious devotee of the art. Even in the profession of radio engineering these are grades through which the studious and the ambitious may progress. An engineer's first-hand, exact knowledge of radio transmission phenomena may well be regarded as a test of his fitness to occupy a place in an advanced grade.

This year the Institute awards the Liebmann prize to an engineer whose researches and investigations into the more difficult element of transmission phenomena have resulted in giving us extensive and useful additions to existing knowledge.

It is my privilege to hand to Dr. Ralph Bown, this award, which goes to him with the respect and admiration of the members of the Board of Direction, and I am sure, with the congratulations of every member of the Institute, both here and abroad.

Introduction of New Officers

The next and final duty which rests upon me is to introduce the president-elect and vice-president elect.

Due largely to the great growth and extension of radio engineering and to the interest taken by the members in the affairs of the Institute, the organization now occupies a place in the family of engineering associations approaching in prestige and importance that enjoyed by the older and larger engineering groups.

The office of president of the Institute of Radio Engineers has become one in which there is great opportunity for constructive doing. The prestige of the Institute as an engineering source is widely recognized. The membership of the Institute is made up mainly of comparatively young men who look to the organization for instruction, for technical information and for opportunity to meet and mingle with their fellows.

The president's task is to initiate and support Institute activities which will insure the availability of these facilities to all members no matter where they may be situated. The gentleman this year elected to occupy the chair is well qualified to carry forward the interests of our beloved Institute to higher levels and to greater usefulness. I take pleasure in presenting president-elect Dr. Ralph Bown.

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RESPONSE OF DR. RALPH BOWN

Incoming President of the Institute of Radio Engineers

I wish to say to you frankly that I am glad to be President of this Institute, that I will do all in my power to be worthy of the responsibility and will carry out the duties of the office to the best of my ability.

I feel sure that in many respects I will be unable to reach the standard of performance exhibited by our retiring President, Mr. McNicol. The devotedness with which he has served the Institute in the past year, while well known to the Board of Directors, is not sufficiently known to the membership at large. To him we owe great advances in the business operation of the Institute. He has devoted probably more than half his time to Institute affairs and has been particularly active in aiding the growth of our various sections. He has just finished presenting to you a report of various activities of the Institute during the past year and some of the plans which have been made for the present year. That report is so complete that I hesitate to burden you with anything further of the same character, so I think that it would perhaps be more fitting if I spent a few minutes in drawing your attention to some other aspect of Institute affairs. I have in mind some reflections which have been engendered by the opportunity I have had during the past year to meet a number of our foreign members. Or perhaps I might better say of our members in foreign countries, since the Institute is so organized that the nationality of a member has no relation to his membership status. In this respect at least, the Institute may be said to be international or world wide in scope.

The total number of foreign members on January 7th was 412 or slightly more than 10 per cent. of our entire membership. Many of these are distinguished engineers of world-wide fame such, for instance, as Senator Marconi, the inventor of wireless telegraphy, Valdemar Poulsen, inventor of the Poulsen Arc, J. Zenneck, the well-known German engineer, E. H. Shaughnessy, Assistant Engineer-in-Chief of the British General Post Office, C. P. Edwards, Director of Radio Service of the Canadian Government, to mention only a few.

If the foreign membership were represented in this

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meeting today in the same ratio as the domestic membership, they would make a very impressive showing and occupy a goodly allotment of seats.

The greatest concentration of foreign members is in England, there being 235 or more than half the total number. One reason for this is that the Proceedings of the Institute are printed in the English language and they carry technical papers which all radio engineers find to be of interest and value. Perhaps this explains also why Japan is prominently represented by 27 members since the English language is frequently used by Japanese engineers and scientists.

It is not uncommon to hear foreign members characterize the Proceedings as the best thing of the kind available in the world. Many foreign members have become members because they wish to obtain the Proceedings and membership is the most satisfactory way to obtain them. Perhaps I should have used the past tense in these remarks, because there seems already to be growing up in the foreign field a feeling that membership is desirable also because of the prestige and professional standing which it implies. And this something we call "prestige"—this intangible asset which we are beginning to acquire—I believe we should guard jealously and foster wisely not only abroad but at home. It is a young and tender plant which we should nurture and cultivate to vigorous maturity as time goes on, for upon it rests in large measure the power of the Institute to serve the public interest and to benefit its individual members.

In doing this, we must remember that the Institute is distinctly American in principle. It does not aim to obtain prestige by virtue of having its membership restricted to a select few who have passed many tests and have acquired sufficient professional standing to add luster to its rolls—on the contrary, it endeavors to enroll and benefit anyone who is interested in radio engineering. To be sure, its higher grades of membership necessarily represent greater proficiency in the art, but suitable grades are open to any who have a sincere desire to associate themselves with radio engineering work. It is a distinctly democratic organization.

High standing, therefore, must flow from recognized high standards and utility of its output. The actions of its governing board must be wise and sound and free from any

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partisan influence. The reports of its technical committees must be authoritative and valuable to the radio industry and to the public generally. The papers which are read and discussed at its meetings must be well written presentations of important matter of broad interest, avoiding the trivial and self-seeking. And perhaps most important of all, its Proceedings must be maintained at the high standard of the past and improved wherever opportunity offers.

We have grown mightily in numbers in the past few years. Let us see to it that we grow even more in utility and in the dignity of service.

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THE CORRELATION OF RADIO RECEPTION WITH SOLAR ACTIVITY AND TERRESTRIAL MAGNETISM*

By
GREENLEAF W. PICKARD

(Consulting Engineer, The Wireless Specialty Apparatus Company, Boston, Massachusetts.)

(Communication from the International Union of Scientific Radio Telegraphy)

One of the outstanding problems today is the nature and cause of those atmospheric changes which produce such diversified effects as weather, magnetic storms and disturbances of radio reception. The problem is meteorological; if this earth has no atmosphere there could be no weather, on an airless planet there could be no long distance communication at broadcasting and higher frequencies and I think magneticians will agree that the phenomena of terrestrial magnetism would be altered if the atmosphere were removed.

The only known important force which acts upon the atmosphere is the complex radiation and emission from that variable star which we call the sun. Changes in this force are caused in two ways; first by the movements of the earth with respect to the sun, and second by actual variations in solar radiation. If the sun maintained a constant radiation, we should have only to consider the earth's rotation on its axis, which gives us night and day, and its movement in an orbit around the sun, which by the changing angle of the solar rays gives us the seasons. If these movements were the only factors involved, weather, terrestrial magnetism and radio reception would follow the calendar to a far greater extent than our measurements indicate.

But in the scheme of things as they are, we find that weather does not go according to the calendar, nor does radio reception. The visual evidence of sunspots, faculae

*Received by the Editor December 20, 1926.
*Presented before the Convention of the Institute of Radio Engineers, January 10th, 1926.

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and prominences tells us that the sun is periodically disturbed, and measurements of the light and heat received by the earth have shown that this varies in general correspondence with visible changes on the sun's disk. Through the work of Abbott and Clayton[1] definite relations have been established between solar changes and weather, which have already been usefully applied to weather forecasting.

Less definite today is our knowledge of the short wave and corpuscular radiation from the sun, which cause ionization and electrical currents in the atmosphere, and even chemical changes. Our only direct indices of these radiations are such things as disturbances of terrestrial magnetism, atmospheric electricity and radio reception, atlhough over long periods they are highly related to sunspots and other visible changes of the sun's surface. And as radio research has not yet become a pure science, we do not have such systematic records to study as those gathered through the years by astronomical and magnetic observatories.

One purpose of this paper is to emphasize the importance of systematic long-period observations of radio reception; it is not an overstatement to say that the picture we hope to draw of the relations of earth and sun is in material part dependent upon such measurements. But at the present time I cannot present a full account of the correlation between radio reception, solar activity and magnetic disturbances, for the simple reason that I have not yet the right sort and amount of data. This paper is therefore merely a progress report setting forth the results so far obtained, which I believe are now sufficiently definite to be of interest to other workers in this field.

Several times in the past twenty years I have attempted systematic measurements of reception from distant stations, in the hope of finding some correlation with other elements. In 1906 I devised[2] a method of accurately measuring received energy from a distant spark station, and for several months in that year I made almost nightly observations at Amesbury, Massachusetts, of the energy received from station CC at South Welfreet, Massachusetts, distant 145 kilometers and operating at 167 kilocycles. I found that the received

[1]Solar Radiation and Weather, or Forecasting Weather from Observations of the Sun. H. H. Clayton, Smithsonian Miscellaneous Collections, Vol. 77, No. 6, June 20, 1925.
Solar Activity and Long Period Weather Changes. H. H. Clayton, Smithsonian Miscellaneous Collections, Vol. 78, No. 4, September 30, 1926

[2]The Measurement of Received Energy at Wireless Stations. Electrical Review, December 15, 1906.

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energy fluctuated over a wide range from day to day, but there was no apparent correlation between these values and either sunspots, terrestrial magnetism or meteorological elements. I am now satisfied that in this series the principal fluctuations were due to changes in the radiation of the transmitter, which was far from constant in either power input, frequency or antenna current.

In 1909[2] I again began a series of reception measurements in Amesbury from station GB at Glace Bay, Nova Scotia, using the audibility meter method of measuring signal intensity. These measurements were repeated at my suggestion by Messrs. Dolbear and Proctor in 1911[3]. Although this series determined an interesting sunset and dawn effect[3] at the time no correlation was found with other elements. More recently I have correlated the night measurements of this series with van Dijk's daily values for the mean magnetic character of days[4], with the result shown below.

Date		Reception Deviation From Mean	Mean Magnetic Character of Day.
January	30, 1909	1.04	1.7
July	25, "	1.37	0.3
"	27, "	0.63	0.7
March	9, 1911	1.41	0.2
"	11-12, "	1.60	0.1
"	13-14, "	0.92	0.7
"	15-16, "	0.25	0.7
"	17-18, "	0.73	0.1
"	21-22, "	1.15	1.4
"	22-23, "	0.94	1.3

The correlation here is $r=-0.214\pm0.204$. This is of course a mathematical equivalent of the Scotch verdict "not proven", but the negative coefficient of correlation is interesting in view of my later work, in which an inverse relation with solar and magnetic measures was definitely found. Those familiar with correlation work will realize that this series was too short for definite results, as it consisted of only ten terms. I am also doubtful of the constancy of the transmitter from night to night.

[2] "Principles of Wireless Telegraphy" G. W. Pierce, New York, 1910, page 135.

[3] "The Effects of Sunlight on the Transmission of Wireless Signals, by B. L. Dolbear and J. A. Proctor, Electrical World, N. Y., Vol. 58, No. 6, August 5, 1911, pages 321-323.

[3] "The Daylight Effect in Radio Telegraphy, by A. E. Kennelly, Proceedings of the Institute of Radio Engineers, Vol. 1, No. 6, July, 1913.

[4] "The Magnetic Character of the Year, by G. van Dijk, Terrestrial Magnetism and Atmospheric Electricity, Vol. XV, No. 3, page 166, and Vol. XVII, No. 3, page 150.

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In one of those all-too-rare compilations of reception data, Marriott has given [1] a tabulation of the variation in nightly range of station DF at Manhattan Beach, New York, over the period October 12, 1908 to October 15, 1909. According to the records of Cheltenham Observatory, Maryland [2] this interval included 33 magnetic storms, one of these —September 25-26, 1909—being the most severe of which we have any record. An examination of Marriott's Chart No. 3, Fig. 1 will show that no less than 26 of these storms coincided with depressions in the range curve.

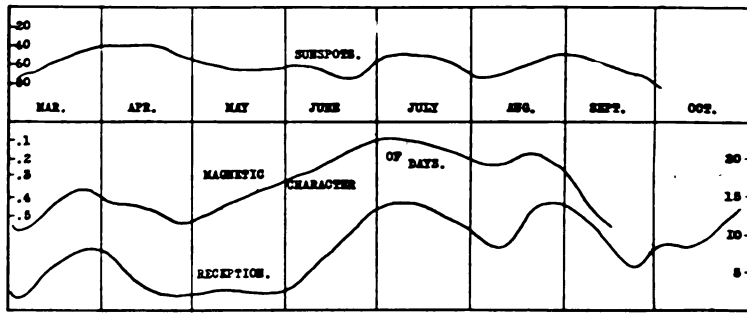

Figure 1.—Moving monthly averages of sunspots, magnetic character of days and reception from WBBM, Chicago. 1926.

Throughout 1909 I kept a reception log at Amesbury, and in this I also find unmistakable coincidences between low reception and magnetic storms. On the evening of September 25 of that year DF was heard but once, with a maximum audibility of 2. On the evening of the 26th DF was not audible at any time, and for several evenings thereafter reception from this station was abnormally low. As I shall show later, our earth recovers its magnetic equilibrium much more rapidly after a magnetic storm than does the coincident depression in radio reception. It is as if some force dealt the earth a blow, simultaneously disturbing its magnetic state and so altering the atmosphere as to lower radio reception. Long after the electrical movements which produce the magnetic storm have subsided, the atmosphere apparently remains unfavorably ionized with respect to radio wave transmission, save at the ultra high frequencies.

[1] Radio Range Variation, by Robert H. Marriott, Proceedings of the Institute of Radio Engineers, Vol. 2, No. 1, March, 1914, pages 37-53.

[2] Terrestrial Magnetism and Atmospheric Electricity, Vol. XIV, No. 1, page 40; No. 2, page 86; No. 3, page 153; No. 4, page 182.

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Beginning with the advent of broadcasting, I recommend once more a systematic measurement of field intensities from distant stations, at first (in 1922) by audibility meter, and later (in 1923) by continuous photographic recording[9]. I again began to notice coincidences between magnetic storms and depressed reception despite the fact that the first years of broadcasting fell in a happy period of minimum solar activity and a magnetically quiescent earth. For convenient reference I give below a list of the magnetic storms of the broadcasting era, with their intensities in gammas. This magnetic unit is 0.00001 gauss, and corres-

Principal Magnetic Storms, 1922-1926 Cheltenham.

Date		Range in Gammas
April	21-22, 1922	186
September	13-15, "	179
October	5- 7, "	193
March	24-25, 1923	183
September	26-28, "	161
November	15-17, "	173
January	29-30, 1924	210
May	21-24, "	187
June	9-11, "	168
June	11-11, "	233
September	7- 8, "	96
September	23-24, "	145
October	23-24, "	119
January	19-20, 1925	103
May	3- 5, "	178
May	30-31, "	91
June	13-14, "	100
June	24-25, "	186
July	25-27, "	122
August	22-23, "	170
September	1- 2, "	153
September	14-15, "	152
September	21-21, "	400
September	24-24, "	186
October	23-24, "	144
November	8- 9, "	129
January	26-27, 1926	548
February	23-25, "	424
March	5- 6, "	286
April	14-17, "	543
May	3- 6, "	147
June	1- 2, "	265
September	9- 9, "	
September	15-15, "	
September	20-21, "	
October	14-16, "	

ponds approximately to one twenty-thousandth of the earth's normal horizontal field at Cheltenham. Although in the ordinary magnetic storm the change in horizontal force is only of the order of one per cent, during the great storm of September 25-26, 1909, there was a change of nearly thirteen per cent in horizontal field.

[9] Short Period Variations in Radio Reception, Proceedings of the Institute of Radio Engineers, Vol. 12, No. 2, April, 1924, pages 119-158.

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A search through newspaper files will show in the radio columns reports of poor broadcast reception coinciding with practically all of the above listed storms. Of this I will give one example. In the Boston Post of September 29, 1923, appeared the following:

> Unusual radio conditions are seriously hampering reception at the present time. This condition was noticeable about this time last year and continued for nearly a week, and has caused trouble for the past two nights. The cause is unknown and while there is hardly a trace of static or fading, it is impossible to tune in stations with any degree of success. While powerful stations like WGY, WMAF and others in New York and Philadelphia may be coaxed in, they seem to cause a click in the phones and disappear, leaving a hum or air rushing sound in the phones. Hundreds of fans listening in during the past two nights have started looking over their sets, changing tubes and batteries and even taking them down. Don't be surprised if this condition continues. The trouble is not with the set but with unusual receiving conditions.

The reference to "about this time last year" is to a period from September 13 to 17, 1922, when the then small but rapidly growing army of broadcast listeners found distant reception poor or impossible. There was a magnetic storm September 13-15, 1922. Similarly, there was another storm September 26-28, 1923, coinciding with the distressful interval so graphically set forth in the quotation above.

Although as I have indicated above, both the technical and popular literature of this art contain sufficient reception data to show a distinct relation to terrestrial magnetism, an accurate knowledge of this and other correlations requires systematic measurements over long periods. An ideal series would consist of daily observations covering the entire radio spectrum over at least a sunspot cycle of eleven years. But this art's solitary boast in such measurements is Austin's series beginning in 1914, of Washington reception from Nauen and other European stations. Unfortunately the low frequency end of the radio spectrum is not particularly sensitive to solar disturbances, and has a distracting relation to certain meteorological elements, so until recently I have not found solar and magnetic relations in this series.

The most sensitive part of the radio spectrum appears to be that band set apart for broadcasting, and more particularly that part from 1000 to 1500 kilocycles. From the viewpoint of this investigation, this is a most fortunate coincidence, for among the radio stations of this world only the broadcasters run "keylocked" for hours each evening, and their unbroken radiation permits a precision of measurement impossible with telegraphically modulated transmitters.

During the fall and winter of 1925 I made an extended

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survey of night reception from distant broadcasting stations, in which I found that in general these stations swung together, a bad night for one being usually a bad night for all. And although I had previously found that short period fading is most violent from stations at moderate distances, it appeared that the greatest differences in mean field from night to night were for stations over a thousand kilometers distant, and operating at the high frequency end of the broadcast band. Finally I selected WBBM at Chicago, operating at 1330 kilocycles, as the best available station for systematic measurements, and early in 1926 I began a series of nightly measurements of this station which have carried through to date without substantial break, and which I hope to continue for at least a year.

The night field from a distant broadcast station is far from uniform, varying in a semi-periodic manner from minute to minute. The amplitude of these fluctuations is so great that no single measurement gives any true idea of the mean field, so it is necessary to either take a large number of separated readings and average them, or, better, to make a continuous record for at least an hour, and then find the mean field by planimeter measurement. Owing to the fact that WBBM shares the frequency of 1330 kilocycles with another Chicago station, WIBO, and to the change in schedule caused by daylight saving time, there is but one hour each evening, 9-10 p. m., 75th meridian time—which can be used throughout the year. With the exception of July, August and September, the field from WBBM has been measured at Newton Centre, Massachusetts. During the summer the measurements were made at Seabrook Beach, New Hampshire, after an overlapping series of observations had shown that on the same evening there was no appreciable difference in mean field at these two points.

As I have already given the method of making such records[9] this will not be described here. An open antenna with an effective height of eight meters was used, with a superheterodyne receiver operating the recorder galvanometer. The record scale was calibrated by introducing a known radio frequency voltage in the antenna circuit, Roberts'[10] ingenious arrangement being used as the source of high frequency.

The nightly field measurements from January to the

[10] Generation and Measurement of Weak Radio-frequency Currents. W. van B. Roberts. Journal Franklin Institute, 201, pages 301-310, March 1926.

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end of November, 1926, are given in an appended table. For nights when WBBM was not operating—Monday night of each week is a silent night for all Chicago stations—or when for any reason I was unable to make observations, I have included an approximate value based upon either reception from a group of other stations and designated by "*e*", or by the arithmetic mean of the values for the nights on each side, indicated by "*m*". In all cases the table gives mean field in microvolts per meter for the period 9-10 p. m., 75th meridian time. This tabulation will perhaps be useful to other investigators, particularly for correlation with solar or meteorological elements.

In the analysis of reception data I have made extensive use of moving averages[11] both for the purpose of showing

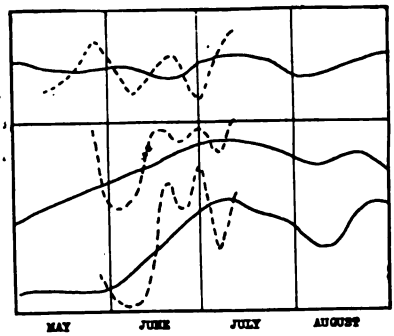

Figure 2.—Sunspots, magnetic character of days and reception, moving monthly averages in broken line. 1926.

general trends over long periods, or of short-time fluctuations. The principal periods which I have used are the week and the month; for all geophysical data this latter period should be made 27 days, in order to eliminate a solar period.

In Fig. 1 the lower curve is a moving monthly average of WBBM reception from March to October, 1926, in microvolts per meter mean field. Above this, with inverted ordinates, is a moving monthly average of magnetic character of day numbers, as given by the Cheltenham Observatory. It is clear that these two curves march together, and a correlation between these elements on a monthly average basis gives $r=-0.89\pm0.06$. This is good correlation, for the

[11] "First Course in Statistical Method" by G. Irving Gavett, New York, 1925, pages 207-209.

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coefficient is high and nearly fifteen times the probable error. The upper curve of this figure is a moving monthly average of the Wolfer Provisional Sunspot Numbers, and has no clear relation to reception or magnetic measures,

Mean field in microvolts per meter at Newton Centre, Mass., from WBBM, Chicago, 9-10 P. M., 1926.

Day	Jan.	Feb.	Mar.	Apr.	May	June	July	Aug.	Sept.	Oct.	Nov.	Dec.
1	....	2.0e	0.0e	3.0	9.1	0.2	39.3	0.2	25.5	9.6	32.0e	0.0e
2	....	8.0e	4.0e	18.2	4.3	4.0e	17.0	0.2m	19.5	7.6	17.3	2.9
3	....	8.0e	2.0e	12.8	8.0e	0.0	3.2	0.2	25.0	8.8	2.8	2.4
4	....	16.0e	0.5	6.6	0.6	0.8	5.1m	6.5	25.0	8.0e	0.5x	2.4
5	....	2.0e	0.2	16.0e	16.0e	0.1	7.0	4.4	16.5	0.3	1.7	7.7
6	....	5.0e	1.0	3.7	0.0	0.8	8.8	0.4	18.3m	7.4	34.0	4.0e
7	....	8.0e	0.0	2.2m	0.2	4.0e	2.9	8.3	20.0	1.3	20.0	18.7
8	....	8.0e	8.0e	0.6	4.0	1.3	17.8	24.6	1.0	8.6	16.0e	6.4
9	....	2.0e	0.3	2.6	4.0	1.0	7.8	19.6m	7.0	35.4	24.0	3.5
10	0.0e	4.0e	3.0	2.8	2.0e	1.7	15.9	14.5	4.8	30.8	10.4	20.6
11	2.0e	2.0e	0.4	3.9	4.0	0.1	24.0	14.2	10.3	30.6m	20.0	13.1e
12	4.0e	0.0e	0.8	16.0e	5.3	0.9	22.9m	13.0	7.5	30.5	25.0	21.6
13	2.0e	0.0e	0.0	3.0	2.8	1.5	21.8	5.7	3.9m	5.6	22.0	16.0e
14	2.0e	0.0e	14.6	0.0	0.0	4.0e	0.7	2.2	0.2	0.0	18.0	6.1
15	2.0e	2.0e	0.0e	0.5	0.0	6.8	7.3m	5.2	3.9m	1.6	16.0e	....
16	1.0e	9.0	0.3	0.2	2.7	8.2	13.8	5.6m	7.5	0.2	7.8	....
17	0.0e	3.8	16.0e	0.0	8.0e	20.2	24.8	6.0	1.9	0.0	6.6	....
18	4.0e	8.3	2.0	0.3	3.3	8.8	10.2	11.9	2.4	4.0e	3.5	....
19	8.0e	2.0	4.0	2.0e	4.8	42.8	19.0m	3.9	2.8	0.0	4.5	....
20	4.0e	3.1	11.0	5.2	5.4	10.8	27.8	0.6	2.2m	0.0	9.4	....
21	16.0e	2.4	4.7	0.4	3.8	8.0e	24.1	14.5	1.6	0.1	7.5	....
22	0.0e	4.0e	8.0e	0.5	0.2	7.5	9.3	17.5	0.0	0.6	4.0e	....
23	4.0e	0.3	28.0	0.1	0.1	17.7	21.7	11.4m	0.0	5.3	20.3	....
24	8.0e	0.4	9.0	0.0	8.0e	16.0e	9.4	5.3	0.1	7.7	2.0	....
25	2.0e	0.2	15.4	9.5	2.5	12.0	13.0	15.5	4.8	8.0e	19.8	....
26	2.0e	0.0	27.0	2.0e	6.1	7.3	11.0m	10.5	6.5	8.6	9.4	....
27	2.0e	0.0	26.0	0.1	2.5	16.9	9.1m	0.5	11.1m	1.9	19.3	....
28	0.0e	0.0	3.2	2.2	7.5	16.0e	7.2	30.5	15.7	6.6	2.0e	....
29	0.0e		2.0e	7.2	4.6	8.0e	6.1	27.5	8.1	9.3	0.0e	....
30	0.0e		1.4	0.0	6.5	12.0	13.2	30.0m	12.1	50.5	4.0e	....
31	0.0e		6.7		8.0e		3.2	32.5		26.7		....
Means.		2.9	3.6	6.4	4.1	4.3	8.0	13.7	11.1	8.8	10.2	12.7

e. Estimated from observations of other stations.
m. Mean of measurements on each side.
x. On November 4 WBBM increased power, and fields from then on are reduced to a basis of 10 amperes in the antenna, which was the current from January 10, to November 4.

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either by inspection or actual correlation on a monthly basis.

But in Fig. 2, which gives in full-line a portion of the three curves of the preceding figure, I have superposed on the moving monthly averages a broken line represently moving weekly averages. Now relation can be seen between sunspots, terrestrial magnetism and reception, as the broken line curves, which represent deviations of weekly averages from monthly means, are obviously similar. An increase of solar activity, indicated by a part of the broken line below the monthly average or full-line axis, is paralleled by an increase in magnetic disturbance, and a decrease in reception.

As we are no longer concerned with the shape of the monthly average curve, this may be flattened out to a zero or base line, and the weekly deviations alone plotted. This

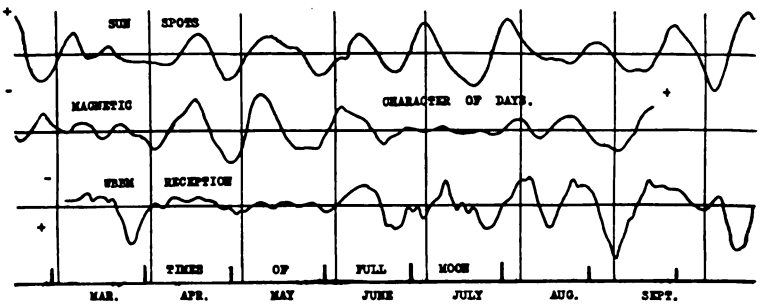

Figure 3.—Weekly departures from monthly means of sunspots, magnetic character of days and reception. 1926.

is done in Fig. 3 for sunspots, magnetic character of days, and reception. Although these three curves do not run absolutely together, there being a marked phase displacement at times, there is the same succession of peaks and hollows; there is no question but that these three elements are related. The period so strongly shown here is 27 days. This is a solar, and not a lunar period, and is in fact due to the recurrent earthward presentation of sunspot or otherwise active areas. More than one casual observer has noted a coincidence between, say, the full of the moon and depressed reception; lunar and solar periods are so nearly alike that once started such coincidences will run through several cycles. But eventually the lunar period will get out of step with reception, and then it will be obvious that

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there is no real relation. This is illustrated in Fig. 3, where I have shown at the bottom the times of full moon. Those who hold lunar superstitions will note that while in August and September the full moon came at or near minimum reception, it came near periods of maximum reception in March, April and May, while in June and July it approximately coincided with normal reception!

Although the rotation of the sun never changes phase, the periodic waxing and waning of the sunspot numbers sometimes do. As the average life of a sunspot is two or three months, a spot or group of spots may persist through several rotations or cycles and then disappear, another

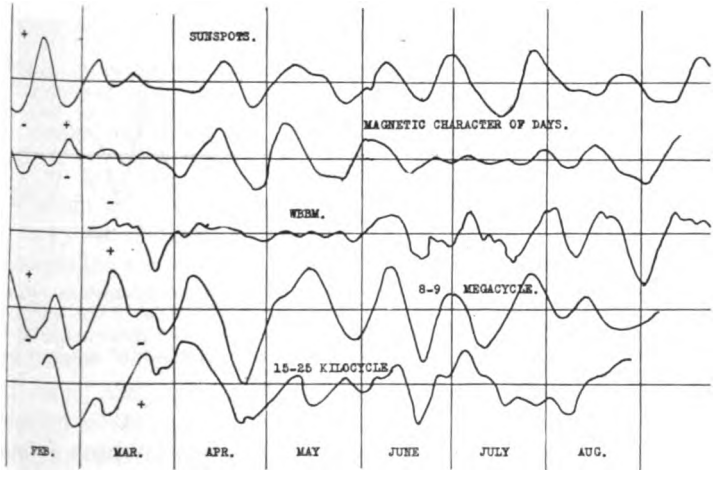

Figure 4.—Weekly averages as deviations from monthly means of sunspots, magnetic character of days, WBBM reception, 8-9 megacycle reception and 15-25 kilocycle reception. 1926.

spot or group of spots appearing at some other point, often in a quite different longitude. This gives a series having a 27-day period, but with occasional changes in phase, much as a bowed violin string vibrates.

Although as I have stated above, the broadcast band is the most sensitive to solar disturbances, effects are in fact produced over the entire radio spectrum. In Fig. 4 I have plotted five elements, sunspots, magnetic character of days, WBBM reception, 8-9 megacycle reception and 15-25 kilocycle reception. The 8-9 megacycle reception is at Washington from NPG, NPM, NBA, NBH, NPL, NIDK and from the U. S. S. Scorpion in the Mediterranean, the trans-

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mission paths being all in darkness. The 15-25 kilocycle reception is also at Washington, from GBR, LPV, FT, AGW, FU, LY, AGS, KET and LPZ, taken at 10.00 a. m., so that the transmission paths were by daylight. The three reception curves are in general alike, the peaks and hollows showing the same succession. But the 8-9 megacycle curve is inverted with respect to the other reception curves, so that this figure gives us the following relation: As solar activity, measured by sunspots, increases, reception in the low-frequency and broadcast bands decreases, but reception at 8-9 megacycles increases. In view of the other inversions which we have found in passing from low to very high-frequency transmission, such as the reversal of the day to

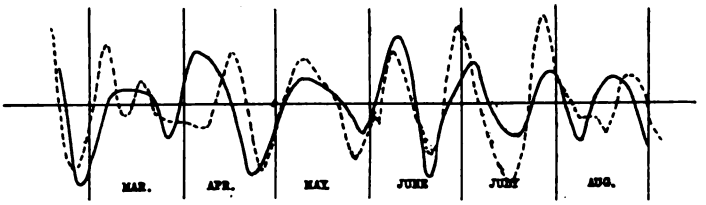

Figure 5.—Average of weekly deviations from monthly means for nine stations in the 12-25 kilocycle band, WBBM at 1330 kilocycles and five stations in 8-9 megacycle band. The broken line is the departure of weekly averages of wolfer provisional sunspot numbers from their monthly means. 1926.

night effects, and the inverted eclipse effect which I found in 1925[12] we are quite prepared to find that these ultra high frequencies, unlike their lower brethren, actually thrive on sunspots and magnetic storms.

Fig. 5 is an arithmetic average of all three reception curves of Fig. 4, with the sunspot curve superposed in broken line. Here is an average of fifteen stations, scattered over nearly a hemisphere, from the Hawaiian Islands, Pacific Coast, the Isthmus, South America, the North Atlantic and points in Europe. And these stations are not only geographically distributed, but they include both ends and the middle of the entire present day radio spectrum. But the averaging of such apparently diverse material has not obliterated the relation with the sun; instead this relation has been accentuated by the process.

[12]The Effect of the Solar Eclipse of January 24, 1925, on Radio Reception. Proceedings of the Institute of Radio Engineers, Vol. 13, No. 5, October, 1925, page 567.

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As I have pointed out above, certain phase differences appear in the various graphs of sunspots, magnetism and reception. Some of these differences are due to the effect which I have already mentioned, the slow recovery of reception from the original disturbance. In Fig. 6 I have plotted

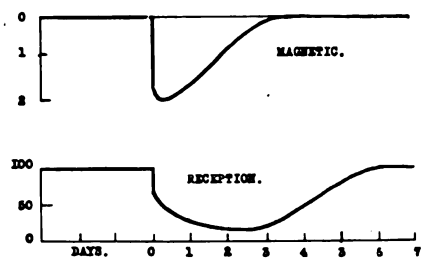

Figure 6.—Magnetic storms and reception. Average of storms of February 23-24, March 5, April 14-16 and May 4-6, 1926. Reception at Newton Centre, Massachusetts, from WBBM, Chicago, Illinois, at 1330 kilocycles, 9-10, 75th meridian time.

an average of four magnetic storms and their accompanying reception depressions, and it will be seen that although the magnetic storm and reception depression begin together, the storm reaches its maximum before reception is at a minimum, and magnetic quiescence returns two or three days before reception is normal.

I have also made a preliminary analysis of WBBM reception with respect to meteorological elements, and particularly with respect to barometric gradients. So far the result has been negative, the field at Newton Centre, when solar and magnetic periods are removed, does not seem to depend in any way upon the relation of the line joining Newton and Chicago to the isobars of the weather map. But there seems to be a slight relation, which I have not yet fully investigated, between barometric activity and reception. Apparently days with great fluctuations of air pressure tend also to be days of low reception. The relation here is probably indirect; that is, barometric activity may be linked with solar disturbances, which are in turn associated with reception.

It is perhaps unlikely that any high correlation between reception and weather elements will be found. Solar disturbances and magnetic storms are world-wide events, whereas weather is rather a local matter. Analysis of weather elements over the whole earth indicate that there

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are areas of positive correlation with sunspots, and also areas of negative correlation. Although I have not yet collected and analyzed reception data from any such collection of receiving points as would fairly represent the earth as a whole, I have found that a bad night for reception in Newton Centre is in general a bad night anywhere in the United States. And I have also found that European reception of distant broadcast stations agrees remarkably well with my measurements of WBBM.

There is some basis for the assumption that reception is principally affected by corpuscular radiation from the sun, perhaps in the form of alpha particles. Maunder has found that magnetic disturbances seem to arise from restricted solar areas, not necessarily including sunspots, and to go out in definite directions, or rather shafts of several degrees diameter, which rotate with the sun. When such a shaft strikes the earth a magnetic storm arises. Such lines of influence are not, she thinks, necessarily radial, but may follow coronal stream lines. I find that, in general, reception is most affected when a spot or group of spots is near the center of the solar disk, that is, when they most nearly face the earth, although there are exceptions. Thus, on November 28-29, 1926, a group of three fair sized spots faced the earth, and coincidently there was a marked depression in reception. Occasionally a large spot may face earthward without any accompanying change in reception, which might be explained either by the assumption that the spot was not accompanied by corpuscular radiation, or on Maunder's hypothesis that its beam was so curved that it did not strike the earth.

But the secrets of this universe yield rather to observation than to pure speculation. When we have a sufficiency of the right kind of data we can frame stable explanations; until then we are groping in the dark. The relation of earth and sun is a dominant one to mankind, and the study of radio transmission phenomena may well throw new light upon this little-understood subject.

I must not conclude without thanking Dr. L. A. Bauer, of the Carnegie Department of Terrestrial Magnetism, for the magnetic and other data which he has supplied, as well as for his many helpful suggestions and also the U. S. Coast and Geodetic Survey for the Cheltenham data which I have so freely used. And I am indebted to Dr. A. H. Taylor of

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the Naval Research Laboratory for the high-frequency reception logs which he so kindly furnished. Finally I must express my appreciation of station WBBM, not only for its involuntary supply of radiation, but for its frequent readings of antenna current each night, which have enabled me to reduce my measurements to a constant basis.

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IMPORTANCE OF LABORATORY MEASUREMENTS IN THE DESIGN OF RADIO RECEIVERS*

W. A. MACDONALD

(Chief Engineer Hazeltine Corporation Laboratory)

The subject of this paper relates primarily to the measured characteristics of radio receivers and the importance of such measurements on the design of commercial broadcast instruments.

It is obvious that an exact knowledge of the individual and over-all characteristics of a radio receiver should be accurately known, yet experience shows that many manufacturers, including some of the largest, are practically unaware of the exact performance of the apparatus they produce.

Some time ago the Hazeltine Corporation was confronted with the problem of measuring the essential characteristics of a large variety of receivers. The first question to be answered was: What are the essential characteristics? Such a question lends itself to considerable discussion, but after careful consideration of all the possible factors it was decided that thirteen fundamental measurements were absolutely necessary in all cases. Additional measurements can be added to meet special circuits or unusual conditions. The thirteen fundamental measurements are as follows:

(1) Voltage step-up of input coupling transformer.
(2) Voltage step-up of 1st tube and coupling transformer.
(3) Voltage step-up of 2nd tube and coupling transformer.
(4) Voltage step-up of following stages.
(5) Complete R. F. amplification from input coup-coil to the detector.
(6) Resonance characteristic of input coupling transformer.

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(7) Resonance characteristic of 1st stage transformer.
(8) Resonance characteristic of 2nd stage transformer.
(9) Resonance characteristic of following stage transformers.
(10 Resonance characteristic of complete R. F. amplifier from input to the detector.
(11) Amplification and frequency characteristics of 1st audio transformer.

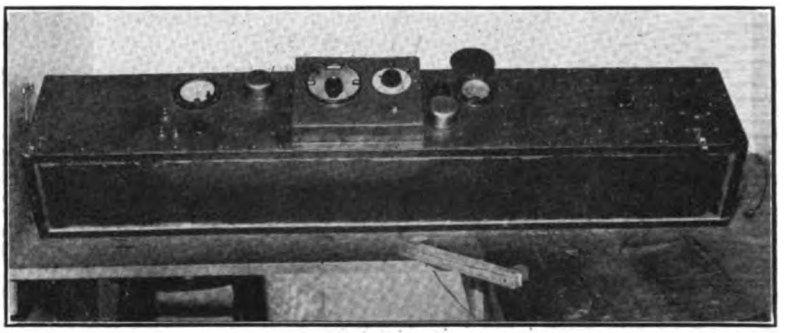

Figure 1

(12) Amplification and frequency characteristics of other audio transformers.
(13) Relative frequency characteristics of complete audio system including detector.

These measurements are selected because they are not

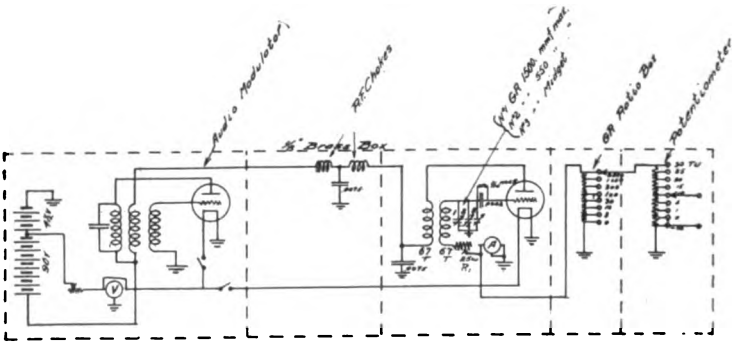

Figure 2—Laboratory Radio Frequency Oscillator

especially difficult to make. They can be readily duplicated and the complete series standardized to a degree where the performance of a radio receiver can be accurately determined.

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The apparatus required is not unduly complicated. It consists essentially of the following:

(1) Precision wave meter.
(2) Radio frequency oscillator.
(3) Audio frequency oscillator.
(4) Vacuum tube voltmeter.

The wave meter can be of any standard make and requires no description.

The radio frequency oscillator may consist of any convenient oscillating circuit, although one giving reasonably

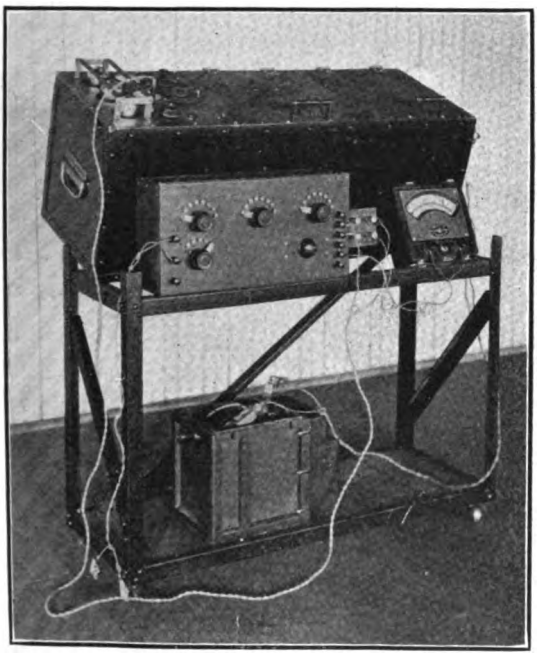

Figure 3

uniform output is preferable. If a simple oscillating circuit is employed it should preferably consist of small inductance and large capacity. Where high accuracy is required the oscillator and all batteries should be completely shielded to eliminate stray fields.

Figure 1 is an illustration of such an oscillator, while Figure 2 is the schematic circuit arrangement. Turning to the illustration, the dials marked $C1$ and $C2$ are the tuning and vernier adjustments. A is the output meter and $R1$ an adjustable resistance for regulating the H. F. output.

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An output voltage attenuator is located at the extreme right.

The audio frequency oscillator may be of a form similar to the radio frequency oscillator. This system might preferably consist of an oscillator and power amplifier and should likewise be shielded.

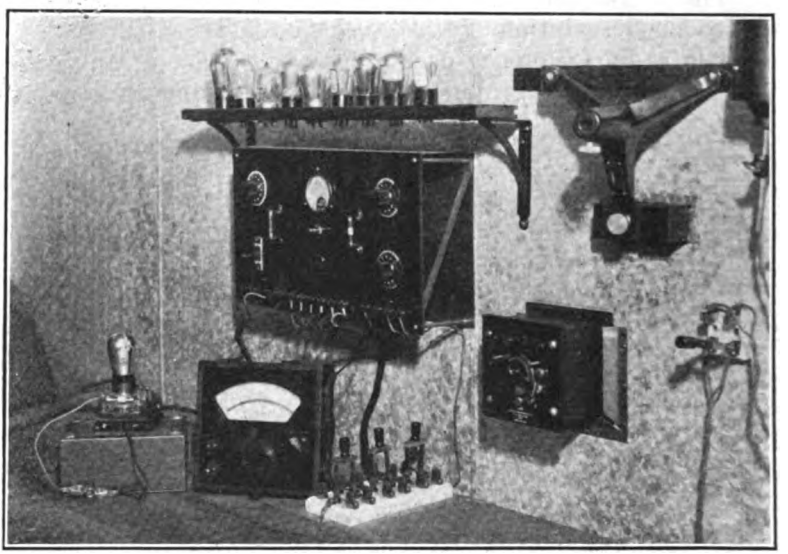

Figure 4

Figure 3 is an illustration of such a system. It is provided with an adjustable frequency range of from 32 to 32000 cycles in octave steps. This oscillator is arranged so that it has a uniform output over the entire frequency range, which of course materially assists in the speed with which measurements can be made.

The vacuum tube voltmeter is similar to that described in the Bell System Technical Journal Vol. III No. 2, Page 185. A UX 112 tube is employed and operated on the square law portion of the curve. This tube should either be debased, or else the indicated results corrected by a suitable constant to compensate for the dielectric loss in the base and socket. The output is read directly on a sensitive micro ammeter or galvanometer. The effect of the steady direct plate current is balanced out of the meter by a reverse current obtained from the heating battery and suitably adjusted by means of resistances.

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Figure 4 is an illustration of a simple form of vacuum tube voltmeter and Figure 5 is the schematic circuit diagram.

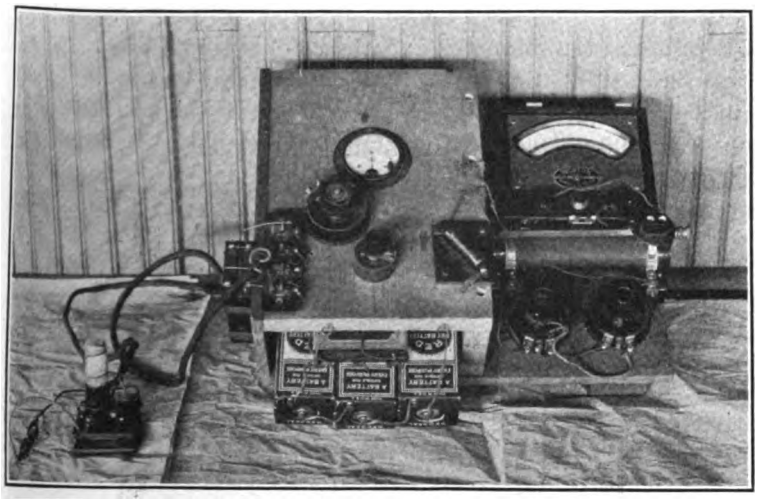

Figure 4A

In order to illustrate the manner in which the measurements are made, a simple laboratory model neutrodyne receiver will be used. It has two stages of tuned radio frequency amplification.

Let us now turn back to the series of measurements previously outlined. No. 1 is a measurement of the voltage

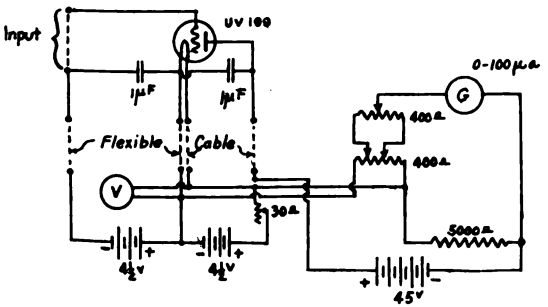

Figure 5—Vacuum Tube Voltmeter

step-up in the input coupling coil. This is made by employing a dummy antenna having suitable characteristics of capacity, inductance and resistance. The dummy antenna

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is connected to the receiver under test. Radio frequency energy is then supplied from the oscillator by means of the output voltage attenuator. The input to the vacuum tube voltmeter is connected across the high voltage terminals of the antenna transformer and the circuit tuned to resonance. The deflection in the output meter of the voltmeter is adjusted to a suitable range by regulation of the voltage attenuator in the output of the oscillator and both the deflection

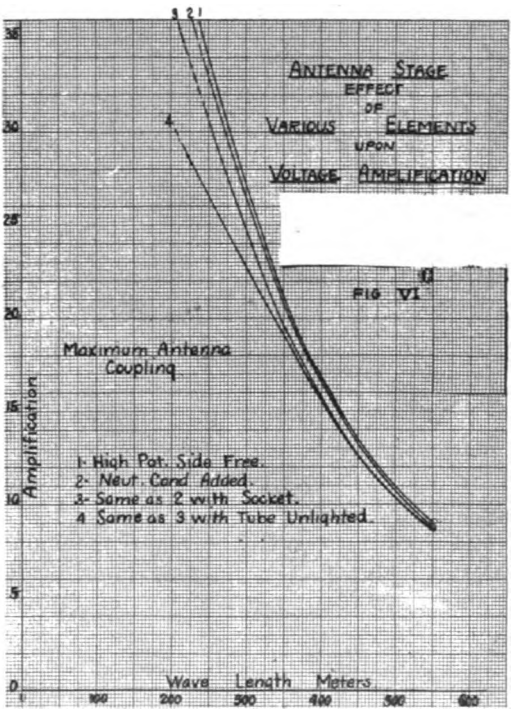

Figure 6

in the meter and value of resistance in the voltage attenuator noted.

The expression for amplification is as follows:

$$A = \sqrt{\frac{\varphi_2}{\varphi_1}} \frac{R_1}{R_2}$$

where φ_2=Deflection of meter
φ_1=Calibration of meter
R_1=Total resistance of voltage attenuator
R_2=Resistance of portion used in measurement.

Figure 6 is a series of curves made on an antenna coup-

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ling transformer and shows the effect of various losses on the amplification characteristics. Curve 1 is the amplification characteristic in which the voltmeter is connected directly across the tuned circuit with other connections removed. Curve 2 is the same measurement with the neutralizing condenser added and shows the reduction in amplification due to the dielectric loss in the condenser. Curves 3 and 4 show additional losses due to the tube socket and tube base.

Figure 7 is a series of curves taken on the first stage of amplification and includes a vacuum tube with its out-

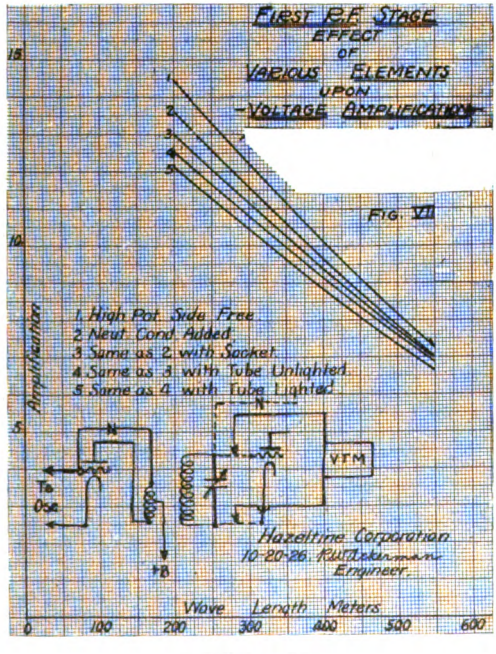

Figure 7

put transformer. In this measurement the dummy antenna is removed and the signal supplied directly to the grid of the tube. Here again the various losses are well defined and consist chiefly of dielectric loss and tube loss.

Figure 8 is the amplification characteristic of the second stage and is made in the same manner as the other measurements. In the case where a grid condenser and leak are employed which impose a load on the tuned circuit, the

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measurement is made with the detector tube in an operative condition.

Before going further it might be interesting if we examined the amplification characteristic obtained from a popular model T. R. F. receiver which depends for its stability, that is freedom from oscillation, on highly damped circuits.

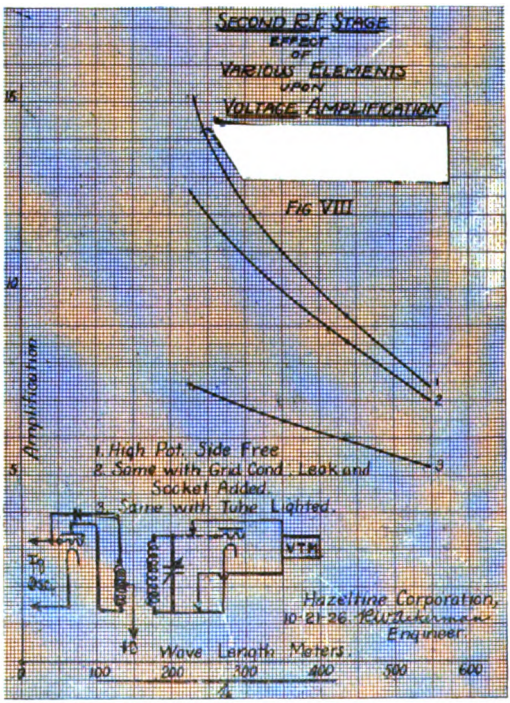

Figure 8

Figure 9 shows the amplification curves obtained from the three tunings of this receiver.

The next point of special interest is the resonance characteristics of the individual tunings. This measurement should be made at two frequencies and can be conveniently done when making the amplification measurements. The curves may be taken by tuning the signal to resonance on the receiver and the detuning the oscillator by means of the vernier adjustment which can be calibrated in kilocycles.

Figure 10 is a series of resonance curves for the various stages of the receiver under test.

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Figure 11 is the resonance curve of the T. R. F. receiver previously mentioned. It is an excellent illustration of the

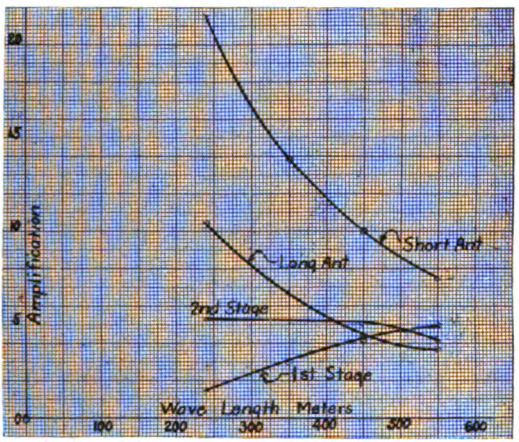

Figure 9—T. R. F. Receiver—R. F. Characteristic

manner in which the frequency admission band of a receiver can be altered by highly damped circuits.

From the measurements so far made the total R. F. am-

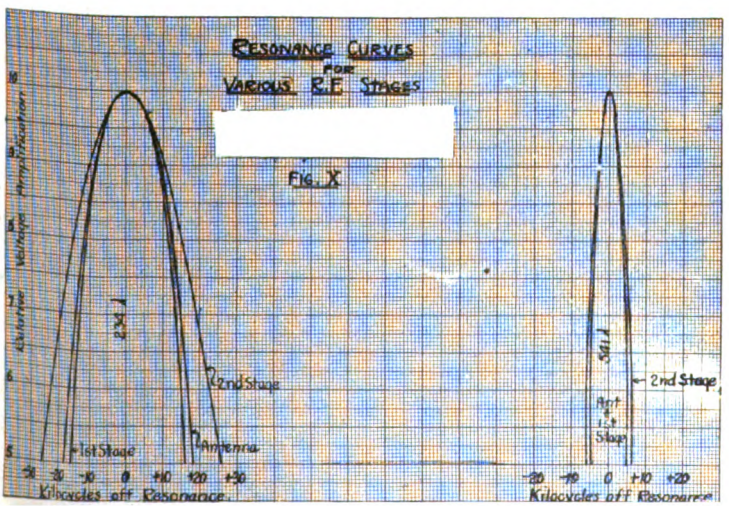

Figure 10

computed, and these are shown in Figures 12 and 13 respectively. Curves of Figures 12 and 13 show the resonance plification as well as the complete resonance curve may be

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characteristic and voltage amplification of the receivers previously referred to.

The next measurements are made on the audio frequency portions of the amplifier. The first involves the voltage step-

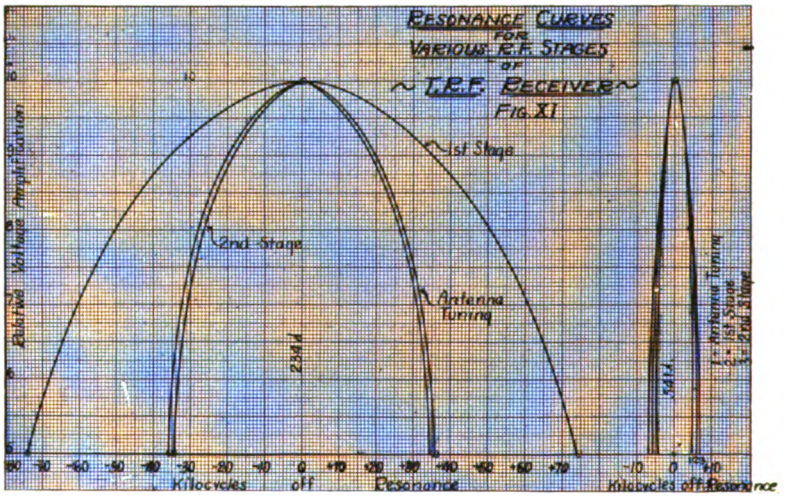

Figure 11

up and frequency characteristic of the individual transformers. This measurement is made by connecting the primary of the transformer to the output voltage divider of the audio

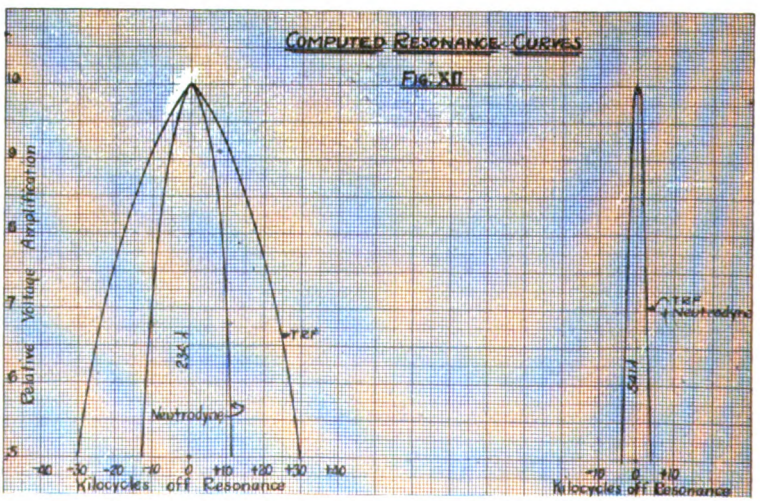

Figure 12

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oscillator. The tube resistance may be simulated by a fixed resistance connected in series with the primary winding. The vacuum tube voltmeter can be connected directly across the secondary terminals of the transformer and the amplification computed as in the R. F. measurements.

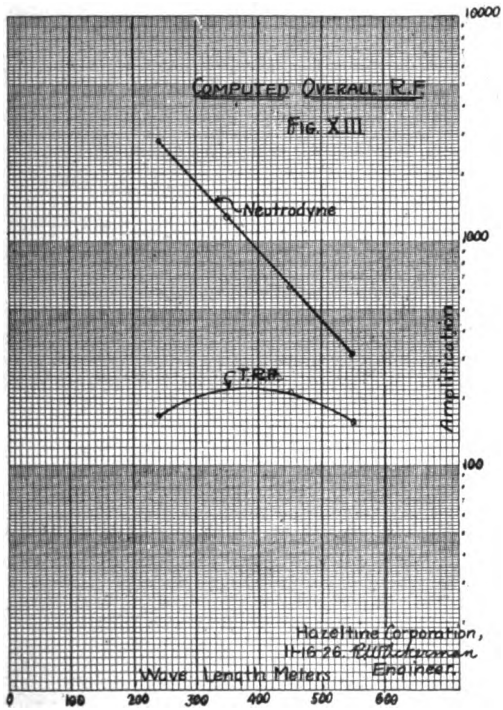

Figure 13

Figure 14 is a typical frequency characteristic of two audio transformers and shows particularly the second rise due to the effect of the leakage inductance. Curve *C* of the same figure shows the computed frequency characteristic of the complete amplifier. Unfortunately, however, this curve has little significance in a complete amplifier because of the effect of the detector grid condenser and the grid leak as well as the regenerative effects due to common battery couplings.

A relative frequency characteristic of the complete amplifier may easily be obtained, however, by supplying the A. F. oscillation to the detector tube in series with the grid leak and measuring the response between grid and filament of the last tube. This will give an amplification curve as

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shown in curve *A*, Figure 15 which is approximately the way a signal is heard in the loudspeaker. With regeneration present, such as due to common "B" batteries, the curve

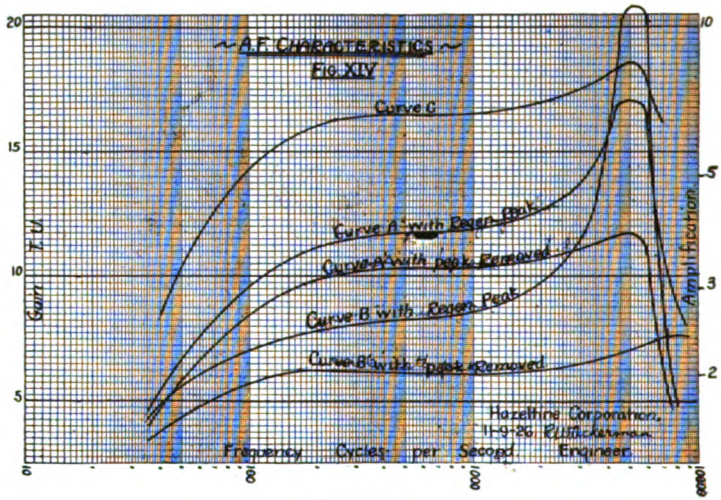

Figure 14

may easily be altered to that shown in curve *B* unless its effect is eliminated by suitable circuit arrangements. Curve *C* of the same figure shows the frequency characteristic of

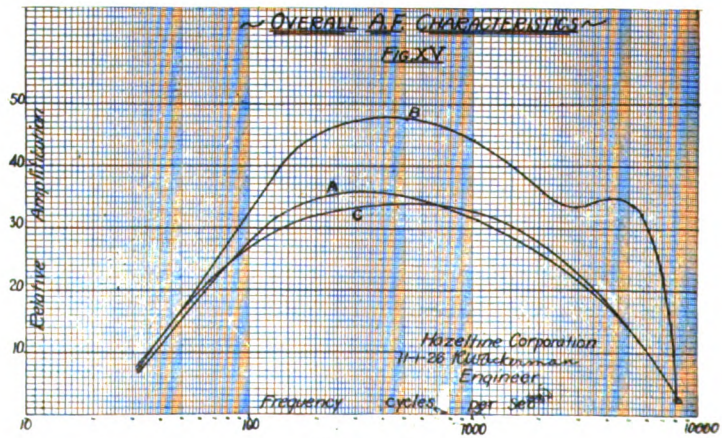

Figure 15

the audio system with the regeneration and second peak removed and is a good example of the present day high class audio system.

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In analyzing the R. F. amplification and resonance curves, it must be remembered that innumerable factors enter into the results which may be obtained. Among these factors are poor transformers, excessive dielectric losses, high resistance of parts, etc. All these effects can be improved and, in many cases, entirely eliminated once their nature has been determined by suitable measurement.

The audio system responds to like treatment and as a rule can be easily corrected when the cause is known.

In conclusion, I wish to acknowledge the assistance of Mr. R. W. Ackerman, a member of the laboratory staff, in the preparation of this paper.

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A THEORETICAL AND EXPERIMENTAL INVESTIGATION OF DETECTION FOR SMALL SIGNALS*

By

E. L. Chaffee and G. H. Browning

Cruft Laboratory, Harvard University

Any electrically conducting device having a non-linear relation between the instantaneous values of output current and impressed voltage will act as a rectifying detector of radio-frequency signals. If the impressed voltage of the signal is a high-frequency e.m.f. of constant amplitude the detector causes a change in the steady value of the output current. If the impressed voltage is a modulated high-frequency e.m.f., *i.e.*, one the amplitude of which varies at a lower frequency, then the output of the detector contains an alternating current of modulation frequency. This process of deriving a current of modulation frequency from a modulated e.m.f. is sometimes known as *demodulation.*

The action of a rectifying detector can be expressed in terms of the impedances of the circuits and the shape of the characteristic curve or curves of the detector. In the case of a crystal detector or a two-electrode vacuum tube, often called a Fleming valve or a *diode*, the theory is very simple. When, however, a three-electrode vacuum tube or triode is used as a detector the complete theory of its action is necessarily complicated because of the larger number of variables. The detecting action can be readily determined experimentally under any prescribed condition, but the theory is of value in showing upon what variables the sensitivity of the detector depends, and also in showing under what conditions and how much distortion results.

Very little has been published on the subject of detection. The definitions of detection proposed in 1919 by Ballantine[1] and by Hazeltine[2] are both incomplete. Carson[3] in 1919 in an article called "A Theoretical Study of the Three-Element Vacuum Tube," gave a very incomplete theory of detection. In 1920,

*Received by the Editor, September 4, 1926.

[1] Proceedings of The Institute of Radio Engineers, volume 7, page 129.

[2] Proceedings of The Institute of Radio Engineers, volume 7, page 172.

[3] Proceedings of The Institute of Radio Engineers, volume 7, page 187.

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G. Breit[4] in a paper called "The Calculation of Detecting and Amplifying Properties of an Electron Tube from its Static Characteristics," presented some very general mathematical expressions for the triode from which some of the theory of detection is derived for unmodulated signals. A second paper by Hulburt and Breit[5] called "The Detecting Efficiency of the Single Electron Tube," gives experimental tests of the formulas derived in Dr. Breit's first paper for certain special cases of detection of unmodulated signals. Mr. Stuart Ballantine, in June 1924, sent to the first author of the present paper some unpublished notes on detection. Although the notes are marked unfinished by Mr. Ballantine, they contain a fairly complete treatment of a special case of grid-circuit rectification. The method of attack adopted by Ballantine is somewhat different from that used in this paper, but the results are in agreement with those here presented for the particular case treated by Ballantine. Since the completion of this paper Mr. Colebrook,[6] of England, has published a series of excellent papers dealing with the mathematical theory of detection for simple detectors and for triodes. Mr. Colebrook's method of treatment is, however, different from that used in this paper and his results differ in form from those presented here.

The present paper[7] presents the theory of detection for both two and three-electrode devices expressed in terms of the circuit impedances and the first and second partial differential coefficients of the static characteristic curves of the device taken at the points on the characteristic curves determined by the steady polarizing voltages. It is then assumed that the impressed signal is so small that for any given steady voltages these coefficients can be assumed constant within the range of the variations due to the signal voltage. A small signal voltage is defined as one less than 0.05 volts r.m.s.

Nomenclature—A brief explanation of the symbols used in this paper will be given at this point.

Small letters in general represent instantaneous values, as e and i for instantaneous e.m.f. and current, respectively. Small letters are also used to denote the slopes of curves which have

[4] "Phys. Rev.," volume XVI, page 387.
[5] "Phys. Rev.," volume XVI, page 408.
[6] "Experimental Wireless," volume II, No. 18, 19, and 20; volume II, No. 26, 27 and volume III, No. 28 and 29.
[7] The theory and method of measuring the detection coefficient given in this paper were worked out by the first author about four years ago and have been presented since that time each year to a class in Vacuum Tubes at Harvard University

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the dimensions of resistance or conductance. For example, if $i=f(e)$ then $k=\frac{di}{de}$ is the conductance for small variations.

Capital letter with a line above represents a steady value of e.m.f. or current, or a resistance for steady current.

Capital letter with a line below represents the maximum amplitude of a sinusoid.

Capital letter with no line above or below denotes the root-mean-square value of an alternating e.m.f. or current, or the value of resistance, inductance, capacity, or impedance for alternating current.

A caret (∧) *over a letter* denotes equivalent value.

A Δ *before a quantity* signifies that it is small but finite.

Subscripts in general denote location of a quantity in a circuit, or frequency, as for instance e_g is the instantaneous value of grid potential, but ω_h is the angular velocity of high frequency.

Bold-face type denotes complex and vector quantities.

I—Theory of Two-Terminal Detector

A—Unmodulated Signal

Detector With No Load. Before taking up the more complex case of the detection by a triode, the simple theory will be presented when the rectifying device is a diode, crystal detector, or electrolytic detector.

In Figure 1, T represents any two-electrode detecting device having a static characteristic given by the relation

$$i=f(e) \tag{1}$$

This function may be expressible only graphically as shown in Figure 2. $\overline{E}_B$ is a constant polarizing potential and $\Delta e_o=\underline{\Delta E}_o$

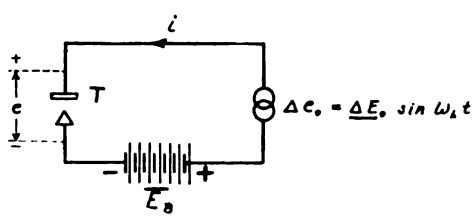

FIGURE 1

$\sin \omega_h t$ is a small impressed radio-frequency potential of frequency $\frac{\omega_h}{2\pi}$.

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When the radio-frequency potential is not impressed, the steady current that flows is

$$\bar{I}=f(\bar{E}_B) \tag{2}$$

Now suppose Δe_o is impressed, then

$$i=f(\bar{E}_B+\underline{\Delta E_o} \sin \omega_h t)$$

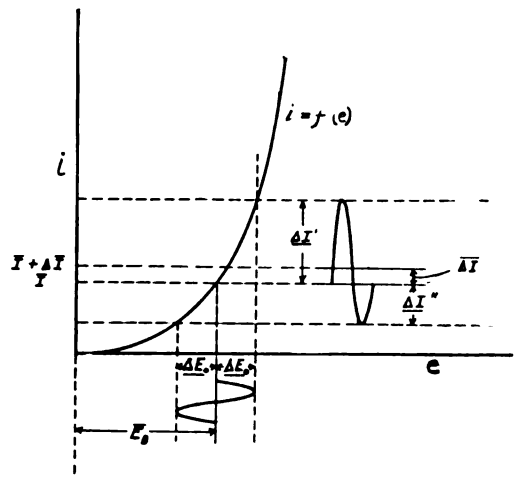

FIGURE 2

The direct current $\bar{I}$ is changed to $\bar{I}+\overline{\Delta I}$, which is the average value of the new current and given by the expression

$$\bar{I}+\overline{\Delta I}=\frac{1}{T}\int_o^T f(\bar{E}_B+\underline{\Delta E_o} \sin \omega_h t)dt. \tag{3}$$

Now $f(\bar{E}_B+\underline{\Delta E_o} \sin \omega_h t)$ can be developed by Taylor's Theorem, giving

$$\bar{I}+\overline{\Delta I}=\frac{1}{T}\int_o^T\left[f(\bar{E}_B)+\frac{df(e)}{de}\underline{\Delta E_o} \sin \omega_h t+\frac{1}{2}\frac{d^2f(e)}{de^2}(\underline{\Delta E_o} \sin \omega_h t)^2+\ldots\right]dt=f(\bar{E}_B)+\frac{1}{4}\frac{d^2i}{de^2}(\underline{\Delta E_o})^2+\ldots \tag{4}$$

Therefore, if terms of higher order than the second are neglected, the change in direct current is given by the expression

$$\overline{\Delta I}=\frac{1}{4}\frac{d^2i}{de^2}(\underline{\Delta E_o})^2 \tag{5}$$

It should be noted that in determining the increase in current, $\overline{\Delta I}$, the first and third order terms in the expansion contribute nothing since they have no average value. The change $\overline{\Delta I}$ is

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then given by the *second order* term. It is, therefore, more accurate to write (5) in the form

$$\overline{\Delta^2 I} = \frac{1}{4}\frac{d^2 i}{d e^2}(\underline{\Delta E_o})^2 = (det.\ I)(\underline{\Delta E_o})^2 \qquad (5\text{-}1)$$

The factor $(det.\ I)$, where

$$(det.\ I) = \frac{1}{4}\frac{d^2 i}{d e^2}\frac{amperes}{(max.\ volts)^2} \qquad (6)$$

is defined as the *current detection coefficient for the device with no load.* If equation (5-I) be multiplied by the total resistance of the circuit for a small increment in current, we have the equivalent steady voltage produced by rectification acting in the output circuit. In the case considered the total resistance is simply $r = \frac{1}{\frac{d i}{d e}}$ of the rectifier. Then the equivalent steady voltage is

$$\overline{\Delta^2 E_o} = r(det.\ I) = \frac{r}{4}\frac{d^2 i}{d e^2}(\underline{\Delta E_o})^2 = \frac{1}{4}\frac{\frac{d^2 i}{d e^2}}{\frac{d i}{d e}}(\underline{\Delta E_o})^2\frac{volts}{(max.\ volts)^2} \qquad (7)$$

The quantity $r\ (det.\ I)$ is the *voltage detection coefficient* of the *rectifier alone* and is denoted by the symbol $(det.\ E)$.

Equation (5-1) gives the rectified current as proportional to the square of the amplitude of the impressed high-frequency alternating e.m.f. Since higher order terms than the second in Taylor's development in equation (4) have been neglected, it is thereby implicitly assumed that any *small* portion of the characteristic curve (equation (1)) of the device about the point determined by $\bar{E}_B$, can be expressed nearly enough by an equation of no higher degree than the second. This, of course, *does not* mean that the curve of equation (1) taken as a whole is expressible as a second degree equation. The assumption above puts no limitation upon the shape of the characteristic curve except that its first and second derivatives shall be everywhere finite.

It is to be noted at this point that we could have obtained the rectified current given in equation (5-1) by the following slightly different procedure. By equation (1) $i = f(e)$.
If i and e have the specific steady values $\bar{I}$ and $\bar{E}_B$, then $\bar{I} = f(\bar{E}_B)$ (see Figure 2). Let $\bar{E}_B$ be given a positive increment $\underline{\Delta E_o}$. As a consequence, $\bar{I}$ will have a positive increment $\underline{\Delta I}'$ such that

$$\bar{I} + \underline{\Delta I}' = f(\bar{E}_B + \underline{\Delta E_o}) \qquad (8)$$

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Now let $\bar{E}_B$ have a negative increment $\underline{\Delta E_o}$ equal to the positive increment, and then $\bar{I}$ will have a negative increment $\underline{\Delta I''}$, such that

$$I - \underline{\Delta I''} = f(\bar{E}_B - \underline{\Delta E_o}) \tag{9}$$

Equations (8) and (9) can be developed by Taylor's Theorem, giving

$$\bar{I} + \underline{\Delta I'} = f(\bar{E}_B) + \frac{df(e)}{de}\underline{\Delta E_o} + \frac{1}{2}\frac{d^2 f(e)}{de^2}(\underline{\Delta E_o})^2 + \ldots \tag{10}$$

$$\bar{I} - \underline{\Delta I''} = f(\bar{E}_B) - \frac{df(e)}{de}\underline{\Delta E_o} + \frac{1}{2}\frac{d^2 f(e)}{de^2}(\underline{\Delta E_o})^2 - \ldots \tag{11}$$

Adding equations (10) and (11) gives

$$\underline{\Delta I'} - \underline{\Delta I''} = \frac{d^2 f(e)}{de^2}(\underline{\Delta E_o})^2 = \frac{d^2 i}{de^2}(\underline{\Delta E_o})^2 \tag{12}$$

The difference $\underline{\Delta I'} - \underline{\Delta I''}$ is a quantity of the second order and must be proportional to the average current $\overline{\Delta^2 I}$ that is produced by the rectifying action of the device. Comparing equations (5-1) and (12) it is evident that *if the impressed potential is sinusoidal*, then

$$\frac{\underline{\Delta I'} - \underline{\Delta I''}}{4} = \overline{\Delta^2 I} \tag{13}$$

Equation (13) shows that the rectified current can be obtained from the two series developments of (10) and (11) instead of by integrating the series as is done in equation (4). This alternative method of obtaining $\overline{\Delta^2 I}$ will be made use of later.

Two-Terminal Detector in Series with Impedance. We may now consider a more general case when, as shown in Figure 3, an impedance Z is included in the circuit. This im-

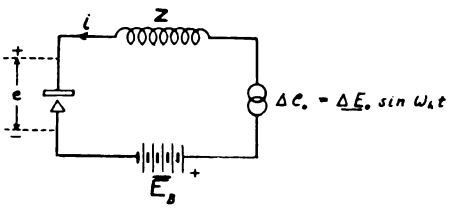

FIGURE 3

pedance Z may have a value $\bar{Z}$ for direct currents and a different value Z_h for a frequency $\frac{\omega_h}{2\pi}$. The detector has for small variations of potential a resistance of

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$$r = \frac{1}{\frac{d f(e)}{d e}} = \frac{1}{\frac{d i}{d e}} \tag{14}$$

Referring to Figure 3, it is obvious that when Δe_o is impressed a current of radio frequency $\frac{\omega_h}{2\pi}$ flows through the circuit, and in addition the steady current is altered by an amount to be denoted by $\overline{\Delta^2 I}$. When no alternating potential is impressed $\overline{I} = f(\overline{E})$, where $\overline{E}$ is the steady potential across the rectifier and different from $\overline{E}_B$. Refer now to Figure 4. The characteristic curve for the rectifier is shown as $i = f(e)$. From point $\overline{E}_B$, a point laid off

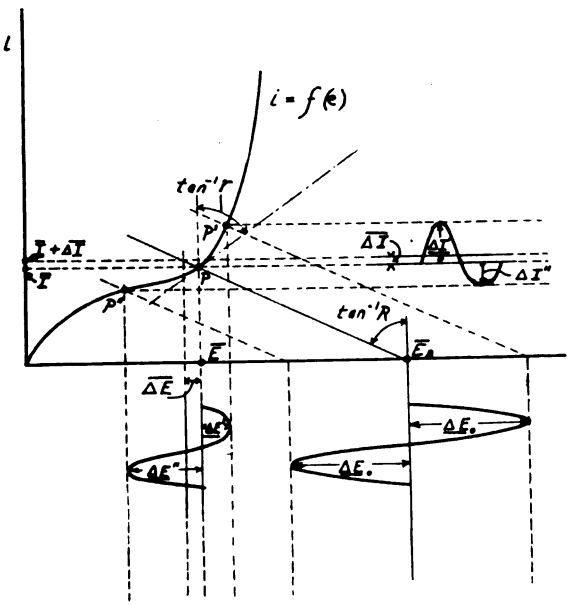

FIGURE 4

on the voltage axis equal to the steady polarizing-battery potential, a line is drawn making an angle with the vertical, the tangent of which is equal to the resistance $\overline{R}$ of Z to steady currents. (For simplicity in representation by diagram, Z is taken as a pure resistance, the same for steady and alternating currents, but the mathematical derivation does not require this limitation.) The point P of intersection of this resistance line and the curve gives the values of $\bar{I}$ and $\bar{E}$. If now $\Delta e_o = \Delta E_o \sin \omega_h t$ is superposed upon $\bar{E}_B$, the new intersection points given by lines drawn parallel to the resistance line from points representing the extreme

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applied voltages $\bar{E}_B+\underline{\Delta E_o}$ and $\bar{E}_B-\underline{\Delta E_o}$ give the extreme currents $\bar{I}+\underline{\Delta I'}$ and $\bar{I}-\underline{\Delta I''}$, and the increments of voltage on the rectifier, *viz.*, $\underline{\Delta E'}$ and $\underline{\Delta E''}$. We can now write

$$\bar{I}+\underline{\Delta I'}=f(\bar{E}+\underline{\Delta E'})$$
$$=f(\bar{E})+\frac{d f(e)}{d e}\underline{\Delta E'}+\frac{1}{2}\frac{d^2 f(e)}{d e^2}(\underline{\Delta E'})^2+\ldots \quad (15)$$

and

$$\bar{I}-\underline{\Delta I''}=f(\bar{E}-\underline{\Delta E''})$$
$$=f(\bar{E})-\frac{d f(e)}{d e}\underline{\Delta E''}+\frac{1}{2}\frac{d^2 f(e)}{d e^2}(\underline{\Delta E''})^2-\ldots \quad (16)$$

Adding (15) and (16), we have

$$\underline{\Delta I'}-\underline{\Delta I''}=\frac{d f(e)}{d e}(\underline{\Delta E'}-\underline{\Delta E''})+$$
$$\frac{d^2 f(e)}{d e^2}\left(\frac{(\underline{\Delta E'})^2+(\underline{\Delta E''})^2}{2}\right)+\ldots \quad (17)$$

Each term of (17) is of the second order in magnitude. By analogy with (13), we can express the differences of increments as steady components of current and voltage. The last term of (17) involves the average of the squares of the increments of voltage and is represented by $(\underline{\Delta E})^2$. Equation (17) can then be written

$$\overline{\Delta^2 I}=\frac{d i}{d e}\overline{\Delta^2 E}+\frac{1}{4}\frac{d^2 i}{d e^2}(\underline{\Delta E})^2+\ldots\ldots \quad (18)$$

$\overline{\Delta^2 E}$ is a steady component of voltage across the rectifier and is equal to $-\bar{R}\,\overline{\Delta^2 I}$. Further, $\underline{\Delta E}$ is the peak radio voltage across the detector and is related to $\underline{\Delta E_o}$ by the obvious relation

$$\underline{\Delta E}=\frac{r}{\sqrt{(r+R_h)^2+X_h^2}}\underline{\Delta E_o} \quad (18\text{-}1)$$

where R_h and X_h are, respectively, the resistance and reactance of the series load Z at the frequency $\frac{\omega_h}{2\pi}$. Then (18) can be written

$$\overline{\Delta^2 I}=-\frac{\bar{R}\,\overline{\Delta^2 I}}{r}+\frac{r^2}{4[(r+R_h)^2+X_h^2]}\cdot\frac{d^2 i}{d e^2}(\underline{\Delta E_o})^2 \quad (19)$$

or

$$\overline{\Delta^2 I}=\frac{r}{r+\bar{R}}\cdot\frac{r^2}{[(r+R_h)^2+X_h^2]}\cdot\frac{1}{4}\cdot\frac{d^2 i}{d e^2}(\underline{\Delta E_o})^2$$
$$=(Det.\ I)(\underline{\Delta E_o})^2 \quad (20)$$

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Hence the *complete current detection coefficient* for the system is

$$(Det.\ I) = \frac{r}{r+\overline{R}} \cdot \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4}\frac{d^2 i}{d e^2} \tag{21}$$

Capital D is used to distinguish the complete detection coefficient from the detection coefficient for the detector alone signified by small d in $(det.\ I)$.

The equivalent voltage of rectification acting in the circuit is

$$\overline{\Delta^2 E} = \frac{r^3}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E_o})^2, \tag{22}$$

and the *equivalent voltage detection coefficient* for this case is

$$(Det.\ E) = \frac{r^3}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4}\frac{d^2 i}{d e^2} \tag{22-1}$$

It is evident from equation (21) and (22-1) that the complete detection coefficient is greater the smaller the radio resistance and reactance, R_h and X_h. If R_h and X_h are negligible, then

$$\overline{\Delta^2 I} = \frac{r}{r+\overline{R}} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E_o})^2 \tag{22-2}$$

and the rectified power is

$$(\overline{\Delta^2 I})^2 \overline{R} = \frac{r^2 \overline{R}}{16(r+\overline{R})^2} \cdot \left(\frac{d^2 i}{d e^2}\right)^2 (\underline{\Delta E_o})^4 \tag{23}$$

The value of R which makes this a maximum is

$$\overline{R} = r \tag{24}$$

This stated in words is as follows:—If there is no impedance offered to the alternating current to be rectified other than that of the rectifier itself, then the maximum rectified power for small values of the impressed e.m.f. is obtained when the resistance of the load or absorbing device for the rectified current is equal to r, the variational resistance of the rectifier.

If the series load is a pure resistance which has the same value for both alternating and direct currents, then the rectified power is

$$(\overline{\Delta^2 I})^2 R = \frac{r^6 R}{16(r+R)^6}\left(\frac{d^2 i}{d e^2}\right)(\underline{\Delta E_o})^4 \tag{25}$$

Solving equation (25) to find the value of R to give maximum power gives

$$R = \frac{r}{5} \tag{26}$$

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B—Detection of Modulated Signal

Case 1. Series Load a Pure Resistance to the Modulation Frequency. In radio communication the detector usually rectifies a modulated radio-frequency e.m.f. giving rise to a current having a frequency equal to the modulation frequency. Figure 5 rep-

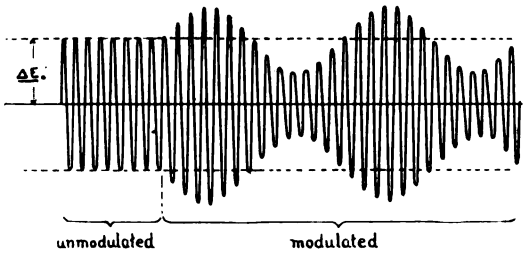

FIGURE 5

resents a sinuosoidally modulated radio-frequency e.m.f., which is expressed mathematically as follows:

$$\Delta e_o = \underline{\Delta E}_o(1+m \sin \omega_l t)\sin \omega_h t \qquad (27)$$

In this expression $\underline{\Delta E}_o$ is the average amplitude of the alternating e.m.f. or the amplitude of the unmodulated radio-frequency e.m.f. of angular velocity ω_h; ω_l is the angular velocity corresponding to the modulation frequency; and m is a factor ranging from 0 for no modulation to 1 for complete modulation, and is a factor giving the fraction of complete modulation and known as the *degree of modulation.*

Suppose now that the modulated e.m.f. of expression (27) is impressed on a rectifying system containing a pure resistance load and having a current detection coefficient (*Det. I*). Then in place of the $(\underline{\Delta E}_o)^2$ of equation (20) must be substituted the square of the coefficient of $\sin \omega_h t$ in equation (27), giving

$$\begin{aligned}\Delta^2 i &= (Det.\ I)(\underline{\Delta E}_o)^2(1+m \sin \omega_l t)^2 \\ &= (Det.\ I)(\underline{\Delta E}_o)^2(1+2m \sin \omega_l t+m^2 \sin^2 \omega_l t) \\ &= (Det.\ I)(\underline{\Delta E}_o)^2\left(1+\frac{m^2}{2}+2m \sin \omega_l t-\frac{m^2}{2}\cos 2\omega_l t\right) \qquad (28)\end{aligned}$$

In the above expression the $\overline{\Delta^2 I}$ of equation (20) is changed to $\Delta^2 i$, to signify that now the rectified component is periodic and has an instantaneous value $\Delta^2 i$. Equation (28) shows that when a modulated e.m.f. is rectified by a system having a curved characteristic and terms of higher order than the second are neglected, three rectified components are obtained as follows:

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1. *Direct current* $= \overline{\Delta^2 I} = \left(1 + \frac{m^2}{2}\right)(Det.\ I)(\underline{\Delta E_o})^2$ (29)

2. *Current of modulation frequency having effective value*

$$= \Delta^2 I_l = \sqrt{2}\, m(Det.\ I)(\underline{\Delta E_o})^2 \quad (30)$$

3. *Current of double modulation frequency having effective value*

$$= \Delta^2 I_{2l} = \frac{m^2}{2\sqrt{2}}(Det.\ I)(\underline{\Delta E_o})^2 \quad (31)$$

In the above special case the complete detection coefficient is the same in all three expressions, 29, 30, and 31. This will not be the case when the load is other than a pure resistance.

It may now be assumed that instead of but one frequency of modulation, there are *two*, as denoted by the following equation which replaces (27).

$$\Delta e_o = \underline{\Delta E_o}(1 + m' \sin \omega_{l'} t + m'' \sin \omega_{l''} t) \sin \omega_h t \quad (31\text{-}1)$$

The resulting current in a rectifying system having a complete detection coefficient of (*Det. I*), due to this doubly-modulated high-frequency potential of (31-1) contains the following components as is evident on substituting the square of the amplitude of $\sin \omega_h t$ of (31-1) in equation (20).

1. *Direct current*

$$= \overline{\Delta^2 I} = \left(1 + \frac{m'^2}{2} + \frac{m''^2}{2}\right)(Det.\ I)(\underline{\Delta E_o})^2 \quad (31\text{-}2)$$

2. *Current of first modulation frequency of effective value*

$$= \Delta^2 I_l' = \sqrt{2}\, m'(Det.\ I)(\underline{\Delta E_o})^2 \quad (31\text{-}3)$$

3. *Current of second modulation frequency of effective value*

$$= \Delta^2 I_{l''} = \sqrt{2}\, m''(Det.\ I)(\underline{\Delta E_o})^2 \quad (31\text{-}4)$$

4. *Current of double first modulation frequency of effective value*

$$= \Delta^2 I_{2l'} = \frac{m'^2}{2\sqrt{2}}(Det.\ I)(\underline{\Delta E_o})^2 \quad (31\text{-}5)$$

5. *Current of double second modulation frequency of effective value*

$$= \Delta^2 I_{2l''} = \frac{m''^2}{2\sqrt{2}}(Det.\ I)(\underline{\Delta E_o})^2 \quad (31\text{-}6)$$

6. *Current of frequency equal to difference of modulation frequencies of effective value*

$$= \Delta^2 I_{(l'-l'')} = \frac{m'\, m''}{\sqrt{2}}(Det.\ I)(\underline{\Delta E_o})^2 \quad (31\text{-}7)$$

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7. Current of frequency equal to sum of modulation frequencies of effective value

$$= \Delta^2 I_{(l'+l'')} = \frac{m'\, m''}{\sqrt{2}} (Det.\ I)(\underline{\Delta E_o})^2 \qquad (31\text{-}8)$$

Equations (29), (30), and (31) show that a rectifying detector used to demodulate a sinusoidally modulated high-frequency signal inevitably introduces a double modulation frequency component which is small in comparison with the main component of modulation frequency only when the modulation is small.

Equations (31-2) to (31-8) further show that if there is more than one frequency of modulation there are introduced beside the main terms of the modulation frequencies and their second harmonics, currents of frequencies equal to both the sum and difference of each pair of modulation frequencies. These sum and difference tones are extremely undesirable as they are dissonant tones. They can be made small only by making the modulation coefficients small.

Case 2—Series Load Not a Pure Resistance at Modulation Frequency. In the preceding treatment of Case 1 the impedance Z of Figure 3 was restricted to a pure resistance for the audio or rectified current, but was not restricted for the modulated or radio-frequency current. The treatment will now be extended to the case where Z may have reactance for the rectified current. We must return to equations (18) and (18-1), both of which hold with no change for the present case. Substituting (18-1) in (18) gives

$$\overline{\Delta^2 I} = \frac{\overline{\Delta^2 E}}{r} + \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \frac{d^2 i}{de^2} (\underline{\Delta E_o})^2 \qquad (32)$$

Now suppose that $(\underline{\Delta E_o})$ has the significance given to it by equation (27). Then $(\underline{\Delta E_o})$ must be replaced by $\underline{\Delta E_o}\ (1+m\ sin\ \omega_l t)$ in (32), which then becomes

$$\Delta^2 i_l = \frac{\Delta^2 e_l}{r} + \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \frac{d^2 i}{d\, e^2} (\underline{\Delta E_o})^2 (1+m\ sin\ \omega_l t)^2$$

$$= \frac{\Delta^2 e_l}{r} + \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \frac{d^2 i}{d\, e^2} (\underline{\Delta E_o})^2 \left(1+\frac{m^2}{2}+2\, m\ sin\ \omega_l t - \frac{m^2}{2} cos\ 2\ \omega_l t\right) \qquad (33)$$

In equation (33) $\Delta^2 i_l$ is now the total instantaneous value of the rectified current which is made up of the constant-current component and the currents of frequencies $\frac{\omega_l}{2\pi}$ and $\frac{2\ \omega_l}{2\pi}$. $\Delta^2 e_l$ is the

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total instantaneous rectified voltage across the rectifying device, and also comprises the instantaneous values of the same three rectified conponents as for $\Delta^2 i_l$. Equation (33) can then be broken up into three equations, the first giving the value of the constant portion of the rectified current, the second the maximum amplitude of the rectified current of modulation frequency $\frac{\omega_l}{2\pi}$, and the third the maximum amplitude of the rectified current of double modulation frequency. These three equations follow:

$$\overline{\Delta^2 I} = \frac{\overline{\Delta^2 E}}{r} + \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E_o})^2 \left(1+\frac{m^2}{2}\right) \tag{34}$$

$$\underline{\Delta^2 I_l} \sin(\omega_l t + \alpha_l) = \frac{\underline{\Delta^2 E_l} \sin(\omega_l t + \beta_l)}{r} + \frac{r}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E_o})^2 \, 2m \sin \omega_l t \tag{35}$$

$$\underline{\Delta^2 I_{2l}} \cos(2\omega_l t + \alpha_{2l}) = \frac{\underline{\Delta^2 E_{2l}} \cos(2\omega_l t + \beta_{2l})}{r} + \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E_o})^2 \frac{m^2}{2} \cos 2\omega_l t \tag{36}$$

Here the phase angles α and β are determined by r and the impedance of the load.

Since the voltage across the rectifying device is the negative of the voltage across the impedance Z, equations (34), (35), and (36) can be rewritten, the last two being expressed in complex notation for root-mean-square values. Complex quantities are signified by bold-face type.

$$\overline{\Delta^2 I} = \frac{r}{r+\overline{R}} \cdot \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} \cdot \left(1+\frac{m^2}{2}\right) (\underline{\Delta E_o})^2 = (\overline{Det.\, I}) \left(1+\frac{m^2}{2}\right) (\underline{\Delta E_o})^2 \tag{37}$$

$$\Delta^2 \boldsymbol{I}_l = \frac{r}{r+\boldsymbol{Z}_l} \cdot \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \frac{d^2 i}{d e^2} \cdot \sqrt{2}\, m\, (\underline{\Delta E_o})^2 = (\boldsymbol{Det.\, I})_l \sqrt{2}\, m (\underline{\Delta E_o})^2 \tag{38}$$

$$\Delta^2 \boldsymbol{I}_{2l} = \frac{r}{r+\boldsymbol{Z}_{2l}} \cdot \frac{r^2}{(r+R_h)^2+X_h^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} \cdot \frac{m}{2\sqrt{2}} (\underline{\Delta E_o})^2 = (\boldsymbol{Det.\, I})_{2l} \frac{m}{2\sqrt{2}} (\underline{\Delta E_o})^2 \tag{39}$$

Obviously in the above equation $(\boldsymbol{Det.\, I})_l$ and $(\boldsymbol{Det.\, I})_{2l}$ are complex quantities and different because of the difference in

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frequency. Considerable advantage arises in expressing the equivalent voltage-detection coefficient, for as will be seen in this case, this quantity is *independent of frequency* and *is not a complex quantity*. Equations (37), (38), and (39) are rewritten, expressed in terms of equivalent rectified voltage, as follows:

$$\hat{\Delta^2 E} = (r+\overline{R})\overline{\Delta^2 I} = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\cdot\frac{d^2 i}{d e^2}\left(1+\frac{m^2}{2}\right)(\underline{\Delta E_o})^2$$

$$= (Det.\ E)\left(1+\frac{m^2}{2}\right)(\underline{\Delta E_o})^2 \qquad (37\text{-}1)$$

$$\hat{\Delta^2} E_l = (r+Z_l)\Delta^2 I_l = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d e^2}\sqrt{2}\, m(\underline{\Delta E_o})^2$$

$$= (Det.\ E)\sqrt{2}\, m(\underline{\Delta E_o})^2 \qquad (38\text{-}1)$$

$$\hat{\Delta^2}\, E_{2l} = (r+Z_{2l})\Delta^2 I_{2l} = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d e^2}\cdot\frac{m^2}{2\sqrt{2}}(\underline{\Delta E_o})^2$$

$$= (Det.\ E)\frac{m^2}{2\sqrt{2}}(\underline{\Delta E_o})^2 \qquad (39\text{-}1)$$

In the above expressions (*Det. E*) is the same for all and is a real quantity given by the following equation:

$$(Det.\ E) = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d e^2} \qquad (40)$$

The complex e.m.fs. of (38-1) and (39-1) do not actually exist, but when used as though they exist give the correct actual values of the rectified currents. The phase of the e.m.f. given by (38-1) is the same as that of the modulation envelope or the modulating e.m.f. The phase of the e.m.f. given by (39-1) is the same as the phase of the double-frequency term in the square of equation (27).

It is obvious that the theory of this case can be extended as in Case 1 to give the magnitude of the difference and sum components where there are two or more frequencies of modulation.

II—Theory of Detection of a Triode

The method used above may now be extended to the three-electrode vacuum tube used as a detector. The connections are shown in Figure 6.

The instantaneous plate current i_p is a function of the instantaneous plate potential e_p and the instantaneous grid potential e_g, and is expressed by the functional equation:

$$i_p = \phi(e_p, e_g) \qquad (41)$$

Similarly the grid current i_g is dependent upon the instantaneous grid potential and to a slight degree upon e_p, but to a fair degree

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of approximation, particularly if the grid is not polarized greatly positively, the grid current can be expressed as a function only of e_g, thus:

$$i_g = \psi(e_g) \tag{42}$$

Unmodulated Signal. We will at first assume that the small unmodulated alternating e.m.f. $\Delta e_o = \underline{\Delta E}_o \sin \omega_h t$ is impressed in the grid circuit of the detector of Figure 6. Because both characteristic equations (41) and (42) are

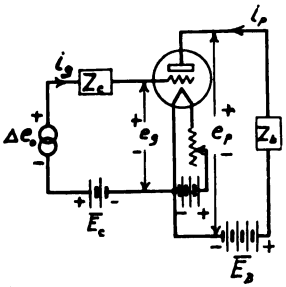

FIGURE 6

curves, the changes of grid and plate voltage, due to the impressed e.m.f. Δe_o, will be made up of steady rectified components as well as variations at frequency $\frac{\omega_h}{2\pi}$. We may then designate the increases of e_g and e_p due to $+\underline{\Delta E}_o$ as $\underline{\Delta E}_g'$ and $\underline{\Delta E}_p'$, respectively, and the decreases in the same quantities as $\underline{\Delta E}_g''$ and $\underline{\Delta E}_p''$, respectively. Using Taylor's Theorem we then have for the increment of plate current,

$$\begin{aligned}
\bar{I}_p + \underline{\Delta I}_p' &= \phi(\bar{E}_p + \underline{\Delta E}_p',\ \bar{E}_g + \underline{\Delta E}_g') \\
&= \phi(\bar{E}_p,\ \bar{E}_g) + \frac{\partial i_p}{\partial e_p}\underline{\Delta E}_p' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p^2}(\underline{\Delta E}_p')^2 + \cdots\cdots \\
&\quad + \frac{\partial i_p}{\partial e_g}\underline{\Delta E}_g' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_g^2}(\underline{\Delta E}_g')^2 + \cdots\cdots \\
&\quad + \frac{\partial^2 i_p}{\partial e_p\, \partial e_g}\underline{\Delta E}_p'\ \underline{\Delta E}_g' + \cdots\cdots
\end{aligned} \tag{43}$$

Similarly the decrease in plate current is given by

$$\begin{aligned}
I_p - \underline{\Delta I}_p'' &= \phi(\bar{E}_p - \underline{\Delta E}_p'',\ \bar{E}_g - \underline{\Delta E}_g'') \\
&= \phi(\bar{E}_p,\ \bar{E}_g) - \frac{\partial i_p}{\partial e_p}\underline{\Delta E}_p'' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p^2}(\underline{\Delta E}_p'')^2 - \cdots\cdots \\
&\quad - \frac{\partial i_p}{\partial e_g}\underline{\Delta E}_g'' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_g^2}(\underline{\Delta E}_g'')^2 - \cdots\cdots \\
&\quad + \frac{\partial^2 i_p}{\partial e_p\, \partial e_g}\underline{\Delta E}_p''\ \underline{\Delta E}_g'' + \cdots\cdots
\end{aligned} \tag{44}$$

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Adding equations (43) and (44) we have

$$\underline{\Delta I_p}' - \underline{\Delta I_p}'' = \frac{\partial i_p}{\partial e_p}(\underline{\Delta E_p}' - \underline{\Delta E_p}'') + \frac{\partial i_p}{\partial e_g}(\underline{\Delta E_g}' - \underline{\Delta E_g}'') + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p^2}[(\underline{\Delta E_p}')^2 + (\underline{\Delta E_p}'')^2] + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_g^2}[(\underline{\Delta E_g}')^2 + (\underline{\Delta E_g}'')^2] + \frac{\partial^2 i_p}{\partial e_p \partial e_g}(\underline{\Delta E_p}'\ \underline{\Delta E_g}' + \underline{\Delta E_p}''\ \underline{\Delta E_g}'') + \cdots\cdots \quad (45)$$

In equations (43), (44), and (45), higher order terms are neglected as being in most cases too small to be important.

Equation (45) can be more simply expressed, remembering that the first three terms are expressible as constant values by equation (13), and that the average of the squares of the increments of the grid and plate voltages may be replaced by single expressions as $(\underline{\Delta E_p})^2 = \frac{(\underline{\Delta E_p}')^2 + (\underline{\Delta E_p}'')^2}{2}$, etc.

$$\overline{\Delta^2 I_p} = \frac{\partial i_p}{\partial e_p}\overline{\Delta^2 E_p} + \frac{\partial i_p}{\partial e_g}\overline{\Delta^2 E_g} + \frac{1}{4}\frac{\partial^2 i_p}{\partial e_p^2}(\underline{\Delta E_p})^2 + \frac{1}{4}\frac{\partial^2 i_p}{\partial e_p^2}(\underline{\Delta E_g})^2 + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p \partial e_g}\underline{\Delta E_p}\ \underline{\Delta E_g} + \cdots\cdots \quad (46)$$

A further simplification of expression (46) is afforded by the well-known definitions of the tube coefficients:

$$\frac{\partial i_p}{\partial e_p} = \textit{plate-to-filament variational conductance} = k_p = \frac{1}{r_p} \quad (47)$$

$$\frac{\partial i_p}{\partial e_g} = \textit{grid-plate variational conductance} = \sigma_p. \quad (48)$$

Rewriting (46), we have

$$\overline{\Delta^2 I_p} = \frac{\overline{\Delta^2 E_p}}{r_p} + \sigma_p\,\overline{\Delta^2 E_g} + \frac{1}{4}\frac{\partial k_p}{\partial e_p}(\underline{\Delta E_p})^2 + \frac{1}{4}\frac{\partial \sigma_p}{\partial e_g}(\underline{\Delta E_g})^2 + \frac{1}{2}\frac{\partial \sigma_p}{\partial e_p}\underline{\Delta E_p}\ \underline{\Delta E_g} + \cdots\cdots \quad (49)$$

In equation (49) $\underline{\Delta E_p}$ is the maximum amplitude of the radio-frequency alternating potential between the plate and filament Now the plate current of frequency $\frac{\omega_h}{2\pi}$ is:

$$\underline{\Delta I_p} = \frac{\mu_p\,\underline{\Delta E_g}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}} \quad (50)$$

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where μ_p is the amplification factor of the tube and equal to $-\frac{\partial e_p}{\partial e_g}$, and R_{bh} and X_{bh} are respectively the resistance and reactance of the plate impedance Z_b at frequency $\frac{\omega_h}{2\pi}$. From (50) we can get $\underline{\Delta E}_p$ by multiplying by $-Z_{bh}$, giving

$$\underline{\Delta E}_p = -\frac{\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\cdot\underline{\Delta E}_g \tag{51}$$

Substituting (51) in (49) gives

$$\overline{\Delta^2 I_p} = \frac{\overline{\Delta^2 E_p}}{r_p}+\sigma_p\overline{\Delta^2 E_g}+\left[\frac{1}{4}\frac{\partial\sigma_g}{\partial e_g}+\frac{1}{4}\frac{\mu_p^2(R_{bh}^2+X_{bh}^2)}{(r_p+R_{bh})^2+X_{bh}^2}\cdot\frac{\partial k_p}{\partial e_p}\right.$$
$$\left.-\frac{\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{2\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\cdot\frac{\partial\sigma_p}{\partial e_p}\right](\underline{\Delta E}_g)^2 \tag{52}$$

The values of $\overline{\Delta^2 E}_g$ and $\underline{\Delta E}_g$ depend upon $\underline{\Delta E}_o$ and the characteristics of the grid circuit. Following the same method for the increments in grid current as for the plate circuit we have

$$\bar{I}_g+\underline{\Delta I}_g' = \psi(\bar{E}_g+\underline{\Delta E}_g')$$
$$= \psi(\bar{E}_g)+\frac{\partial i_g}{\partial e_g}\underline{\Delta E}_g'+\frac{1}{2}\frac{\partial^2 i_g}{\partial e_g^2}(\underline{\Delta E}_g')^2+\cdots\cdots \tag{53}$$

and

$$\bar{I}_g-\Delta I_g'' = \psi(\bar{E}_g-\underline{\Delta E}_g'')$$
$$= \psi(\bar{E}_g)-\frac{\partial i_g}{\partial e_g}\underline{\Delta E}_g''+\frac{1}{2}\frac{\partial^2 i_g}{\partial e_g^2}(\underline{\Delta E}_g'')^2+\cdots\cdots \tag{54}$$

Adding (53) and (54) gives

$$\underline{\Delta I}_g'-\underline{\Delta I}_g'' = \frac{\partial i_g}{\partial e_g}(\underline{\Delta E}_g'-\underline{\Delta E}_g'')+\frac{1}{2}\frac{\partial^2 i_g}{\partial e_g^2}[(\underline{\Delta E}_g')^2$$
$$+(\underline{\Delta E}_g'')^2]+\cdots\cdots \tag{55}$$

Converting (55) by equation (13) and replacing $\frac{\partial i_g}{\partial e_g}$ by the grid variational conductance $k_g=\frac{1}{r_g}$, we have

$$\overline{\Delta^2 I}_g = \frac{\overline{\Delta^2 E}_g}{r_g}+\frac{1}{4}\frac{\partial k_g}{\partial e_g}(\underline{\Delta E}_g)^2+\cdots\cdots \tag{56}$$

We now wish to express the maximum amplitude of the alternating e.m.f. $\underline{\Delta E}_g$ between grid and filament in terms of $\underline{\Delta E}_o$, the amplitude of the impressed e.m.f. The amplitude of the high-frequency alternating current flowing in the grid circuit is

$$\underline{\Delta I}_g = \frac{\underline{\Delta E}_o}{\sqrt{(R_{ch}+r_{gh})^2+(X_{ch}+j_{gh})^2}} \tag{57}$$

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where R_{ch} and X_{ch} are respectively the resistance and reactance to frequency $\frac{\omega_h}{2\pi}$ of the series impedance Z_c in the grid circuit, and $\hat{r}_{gh}$ and $\hat{x}_{gh}$ are, respectively, the equivalent grid input resistance and reactance of the vacuum tube at frequency $\frac{\omega_h}{2\pi}$. (We have up to this point neglected any effects caused by capacity between the tube elements. It is well known, however, that the tube capacities, notably the capacity between the grid and plate, give the tube an equivalent input resistance different from r_g and an input reactance which may be either positive or negative, dependent upon the load in the plate circuit.) But

$$\underline{\Delta E}_g = \underline{\Delta I}_g \sqrt{\hat{r}_{gh}^2 + \hat{x}_{gh}^2} = \frac{\sqrt{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}}{\sqrt{(R_{ch}+\hat{r}_{gh})^2 + (X_{ch}+\hat{x}_{gh})^2}} \cdot \underline{\Delta E}_o \qquad (58)$$

Equation (58) is now to be substituted into (52) and (56), and we have

$$\overline{\Delta^2 I_p} = \frac{\overline{\Delta^2 E_p}}{r_p} + \sigma_p \overline{\Delta^2 E_g} + \left[\frac{1}{4}\frac{\partial \sigma_p}{\partial e_g} + \frac{\mu_p^2(R_{bh}^2+X_{bh}^2)}{4[(r_p+R_{bh})^2+X_{bh}^2]} \cdot \frac{\partial k_p}{\partial e_p} - \frac{\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{2\sqrt{(r_p+R_{bh})^2+X_{bh}^2}} \cdot \frac{\partial \sigma_p}{\partial e_p}\right] \frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(R_{ch}+\hat{r}_{gh})^2+(X_{ch}+\hat{x}_{gh})^2} \cdot (\underline{\Delta E}_o)^2 \qquad (59)$$

and

$$\overline{\Delta^2 I_g} = \frac{\overline{\Delta^2 E_g}}{r_g} + \frac{1}{4}\frac{\partial k_g}{\partial e_g} \cdot \frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(R_{ch}+\hat{r}_{gh})^2+(X_{ch}+\hat{x}_{gh})^2} (\underline{\Delta E}_o)^2 \qquad (60)$$

As an abbreviation let the square bracket of (59) be denoted by A

$$A = \frac{1}{4}\frac{\partial \sigma_p}{\partial e_g} + \frac{\mu_p^2(R_{bh}^2+X_{bh}^2)}{4[(r_p+R_{bh})^2+X_{bh}^2]} \cdot \frac{\partial k_p}{\partial e_p} - \frac{\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{2\sqrt{(r_p+R_{bh})^2+X_{bh}^2}} \cdot \frac{\partial \sigma_p}{\partial e_p}, \qquad (61)$$

and let B stand for

$$B = \frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(R_{ch}+\hat{r}_{gh})^2+(X_{ch}+\hat{x}_{gh})^2} \qquad (62)$$

Modulated Signal. So far we have assumed that the impressed alternating e.m.f. Δe_o has a constant amplitude $\underline{\Delta E}_o$. Now, following the method previously adopted, we will suppose that the impressed e.m.f. of frequency $\frac{\omega_h}{2\pi}$ is modulated at

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a frequency $\frac{\omega_l}{2\pi}$ as indicated by equation (27). The maximum amplitude $\underline{\Delta E}_o$, assumed constant in the preceding equations, is now a function of t and the value $\underline{\Delta E}_o$ is to be replaced in equations (59) and (60) by

$$\underline{\Delta E}_o(1+m \sin \omega_l t)$$

Because $\underline{\Delta E}_o$ in these equations is squared, three component terms result, one of constant amplitude, one of frequency $\frac{\omega_l}{2\pi}$, and the third of double this frequency. The quantities $\overline{\Delta^2 I}_p$, $\overline{\Delta^2 I}_g$, $\overline{\Delta^2 E}_p$, $\overline{\Delta^2 E}_g$ must now be considered functions of time and each will be made up of three components. With this in mind we may again rewrite (59) and (60) as follows:

$$\begin{aligned}
&\overline{\Delta^2 I}_p+\underline{\Delta I}_{pl}\sin(\omega_l t+\alpha_l)-\underline{\Delta I}_{p2l}\cos(2\omega_l t+\alpha_{2l}) \\
&\quad=\frac{\overline{\Delta^2 E}_p+\underline{\Delta E}_{pl}\sin(\omega_l t+\beta_l)-\underline{\Delta E}_{p2l}\cos(2\omega_l t+\beta_l)}{r_p} \\
&\quad+\sigma_p(\overline{\Delta^2 E}_g+\underline{\Delta E}_{gl})\sin(\omega_l t+\gamma_l)-\underline{\Delta E}_{g2l}\cos(2\omega_l t+\gamma_{2l}) \\
&\quad+A\,B\left(1+\frac{m^2}{2}+2m\sin\omega_l t-\frac{m^2}{2}\cos 2\omega_l t\right)(\underline{\Delta E}_o)^2 \qquad (63)
\end{aligned}$$

$$\begin{aligned}
&\overline{\Delta^2 I}_g+\underline{\Delta I}_{gl}\sin(\omega_l t+\theta_l)-\underline{\Delta I}_{g2l}\cos(2\omega_l t+\theta_{2l}) \\
&\quad=\frac{\overline{\Delta^2 E}_g+\underline{\Delta E}_{gl}\sin(\omega_l t+\theta_l)-\underline{\Delta E}_{g2l}\cos(2\omega_l t+\theta_{2l})}{r_g} \\
&\quad+\frac{B}{4}\frac{\partial k_g}{\partial e_g}\left(1+\frac{m^2}{2}+2m\sin\omega_l t-\frac{m^2}{2}\cos 2\omega_l t\right)(\underline{\Delta E}_o)^2 \qquad (64)
\end{aligned}$$

Each of equations (61) and (62) now breaks up into three equations, one for each component. If now these component equations be expressed in effective complex values we have:

$$\left.\begin{aligned}
\overline{\Delta^2 I}_p&=\frac{\overline{\Delta^2 E}_p}{r_p}+\sigma_p\overline{\Delta^2 E}_g+A\,B\left(1+\frac{m^2}{2}\right)(\underline{\Delta E}_o)^2 \qquad (63\text{-}1)\\
\overline{\Delta^2 I}_g&=\frac{\overline{\Delta^2 E}_g}{r_g}+\frac{B}{4}\frac{\partial k_g}{\partial e_g}\left(1+\frac{m^2}{2}\right)(\underline{\Delta E}_o)^2 \qquad (63\text{-}2)
\end{aligned}\right\}\ \textit{constant term}$$

$$\left.\begin{aligned}
\Delta^2 I_{pl}&=\frac{\Delta^2 E_{pl}}{r_p}+\sigma_p\Delta^2 E_{gl}+A\,B\sqrt{2}\,m(\underline{\Delta E}_o)^2 \qquad (64\text{-}1)\\
\Delta^2 I_{gl}&=\frac{\Delta^2 E_{gl}}{r_g}+\frac{B}{4}\frac{\partial k_g}{\partial e_g}\sqrt{2}\,m(\underline{\Delta E}_o)^2 \qquad (64\text{-}2)
\end{aligned}\right\}\ \textit{modulation frequency}$$

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$$\Delta^2 I_{p2l} = \frac{\Delta^2 E_{p2l}}{r_p} + \sigma_p \underline{\Delta^2 E_{g2l}} + A\, B \frac{m^2}{2\sqrt{2}} (\underline{\Delta E_o})^2 \qquad \text{(65-1)}$$

$$\Delta^2 I_{g2l} = \frac{\Delta^2 E_{g2l}}{r_g} + \frac{B}{4} \frac{\partial k_g}{\partial e_g} \frac{m^2}{2\sqrt{2}} (\underline{\Delta E_o})^2 \qquad \text{(65-2)}$$

double modulation frequency

The two equations for each component can now be combined and simplified by means of the following relations:

$$\overline{\Delta^2 E_{pl}} = -\overline{R}_b \overline{\Delta^2 E_p} \qquad \overline{\Delta^2 E_g} = -\overline{R}_c \overline{\Delta^2 I_g} \qquad (66)$$

$$\Delta^2 E_{pl} = -Z_{bl} \Delta^2 I_{pl} \qquad \Delta^2 E_{gl} = -Z_{cl} \Delta^2 I_{gl} \qquad (67)$$

$$\Delta^2 E_{p2l} = -Z_{b2l} \Delta^2 I_{p2l} \qquad \Delta^2 E_{g2l} = -Z_{c2l} \Delta^2 I_{g2l} \qquad (68)$$

giving the final expression for the three rectified components as follows:

$$\overline{\Delta^2 I_p} = \frac{r_p}{r_p + \overline{R}_b} \left[-\frac{\sigma_p r_g \overline{R}_c}{4(r_g + \overline{R}_c)} \cdot \frac{\partial k_g}{\partial e_g} + A \right] B \left(1 + \frac{m^2}{2}\right) (\underline{\Delta E_o})^2$$

$$= (\overline{Det.\ I}) \left(1 + \frac{m^2}{2}\right) (\underline{\Delta E_o}^2) \qquad (69)$$

$$\Delta^2 I_{pl} = \frac{r_p}{r_p + Z_{bl}} \left[-\frac{\sigma_p r_g Z_{cl}}{4(r_g + Z_{cl})} \cdot \frac{\partial k_g}{\partial e_g} + A \right] B \sqrt{2}\, m (\underline{\Delta E_o})^2$$

$$= (\boldsymbol{Det.\ I})_l \sqrt{2}\, m (\underline{\Delta E_o})^2 \qquad (70)$$

$$\Delta^2 I_{p2l} = \frac{r_p}{r_p + Z_{b2l}} \left[-\frac{\sigma_p r_g Z_{c2l}}{4(r_g + Z_{c2l})} \cdot \frac{\partial k_g}{\partial e_g} + A \right] B \frac{m^2}{2\sqrt{2}} (\underline{\Delta E_o})^2$$

$$= (\boldsymbol{Det.\ I})_{2l} \frac{m^2}{2\sqrt{2}} (\underline{\Delta E_o})^2 \qquad (71)$$

The values of the equivalent current and voltage detection coefficients are listed below for convenient reference.

$$(\overline{Det.\ I}) = \frac{r_p}{4(r_p + \overline{R}_b)} \left[-\frac{r_g \overline{R}_c}{r_g + \overline{R}_c} \sigma_p \frac{\partial k_g}{\partial e_g} + \frac{\partial \sigma_p}{\partial e_g} + \frac{R_{bh}^2 + X_{bh}^2}{(r_p + R_{bh})^2 + X_{bh}^2} \mu_p^2 \frac{\partial k_p}{\partial e_p} - \frac{2\sqrt{R_{bh}^2 + X_{bh}^2}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}} \mu_p \frac{\partial \sigma_p}{\partial e_p} \right] \left[\frac{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}{(\hat{r}_{gh} + R_{ch})^2 + (\hat{x}_{gh} + X_{ch})^2} \right]$$

$$(\overline{Det.\ E}) = (r_p + \overline{R}_b)(\overline{Det.\ I}) \qquad (72)$$

steady component

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$$
\left.\begin{aligned}
(\textbf{\textit{Det. I}})_1 &= \frac{r_p}{4(r_p+Z_{b1})}\left[-\frac{r_g Z_{c1}}{r_g+Z_{c1}}\sigma_p\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_g}\right.\\
&\quad+\frac{R_{bh}^2+X_{bh}^2}{(r_p+R_{bh})^2+X_{bh}^2}\mu_p{}^2\frac{\partial k_p}{\partial e_p}\\
&\quad\left.-\frac{2\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\mu_p\frac{\partial \sigma_p}{\partial e_p}\right]\\
&\quad\left[\frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(\hat{r}_{gh}+R_{ch})^2+(\hat{x}_{gh}+X_{ch})^2}\right]\\
(\textbf{\textit{Det. E}})_1 &= (r_p+Z_{b1})(\textbf{\textit{Det. I}})_1
\end{aligned}\right\}\ \textit{component of modulation freqency}\qquad(73)
$$

$$
\left.\begin{aligned}
(\textbf{\textit{Det. I}})_{21} &= \frac{r_p}{4(r_p+Z_{b21})}\left[-\frac{r_g Z_{c21}}{r_g+Z_{c21}}\sigma_p\frac{\partial k_g}{\partial e_g}\right.\\
&\quad+\frac{\partial \sigma_p}{\partial e_g}+\frac{R_{bh}^2+X_{bh}^2}{(r_p+R_{bh})^2+X_{bh}^2}\mu_p{}^2\frac{\partial k_p}{\partial e_p}\\
&\quad\left.-\frac{\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\mu_p\frac{\partial \sigma_p}{\partial e_p}\right]\cdot\\
&\quad\left[\frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(\hat{r}_{gh}+R_{ch})^2+(\hat{x}_{gh}+X_{ch})^2}\right]\\
(\textbf{\textit{Det. E}})_{21} &= (r_p+Z_{b21})(\textbf{\textit{Det. I}})_{21}
\end{aligned}\right\}\ \textit{component of double modulation frequency}\qquad(74)
$$

Equations (72), (73), and (74) give the final and complete values, to the approximation allowed by the assumptions made, of the change in plate current and the currents of modulation frequency and of double modulation frequency when the impressed signal is very small. It is evident that the equivalent current and voltage detection coefficients $(\textbf{\textit{Det. I}})_1$ $(\textbf{\textit{Det. I}})_{21}$, $(\textbf{\textit{Det. E}})_1$, and $(\textbf{\textit{Det. E}})_{21}$ are, in the case of the triode, all complex quantities. Attention is called to the fact that the value of $(\textbf{\textit{Det. I}})$ depends upon the load in the plate circuit, whereas the voltage detection coefficient $(\textbf{\textit{Det. E}})$ is independent of the plate load. $(\textbf{\textit{Det. E}})$ gives the equivalent voltage of audio frequency which is considered to be introduced into the plate circuit by detection and from which the current in the plate circuit can be calculated when the total impedance of the plate circuit is known. The voltage detection coefficient is, therefore, a better quantity to express the detecting ability of a triode.

It is evident that the theory can be easily extended to give the amplitudes of the sum and difference components when there is more than one frequency of modulation.

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A very simple transformation makes somewhat easier the interpretation of the equations (69) to (74). Since $\sigma_p = \mu_p k_p$, then

$$\frac{\partial \sigma_p}{\partial e_p} = \mu_p \frac{\partial k_p}{\partial e_p} + k_p \frac{\partial \mu_p}{\partial e_p} \tag{75}$$

This relation can now be substituted in equations (69) to (74). This simplification, however, is expressed only in the case of equation (70), for this equation is the one of greatest interest. The other equations can, by comparison with the following new form for (70), be easily changed.

$$\Delta^2 I_{pl} = \frac{r_p}{4(r_p + Z_{bl})}\left[-\frac{r_g Z_{cl}}{r_g + Z_{cl}} \sigma_p \frac{\partial k_g}{\partial e_g} + \frac{\partial \sigma_p}{\partial e_g} \right.$$
$$- \frac{2\sqrt{R_{bh}^2 + X_{bh}^2}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}} \sigma_p \frac{\partial \mu_p}{\partial e_p} - \frac{\sqrt{R_{bh}^2 + X_{bh}^2}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}}$$
$$\left(2 - \frac{\sqrt{R_{bh}^2 + X_{bh}^2}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}}\right) \mu_p^2 \frac{\partial k_p}{\partial e_p} \Big] \cdot$$
$$\left[\frac{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}{(\hat{r}_{gh} + R_{ch})^2 + (\hat{x}_{gh} + X_{ch})^2}\right] \sqrt{2}\, m(\underline{\Delta E_g})^2 \tag{76}$$

An examination of equation (73) or the transformed form given by equation (76), giving the audio current of modulation frequency, shows that detection or demodulation by a triode is a very complex process depending upon the audio and radio impedances in both the grid and place circuits as well as upon the tube coefficients and several of their derivatives. It is impossible to form any idea of the magnitude of the several terms of the expression without knowing the values of the tube coefficients and their derivatives. For a hard tube the grid-plate conductance σ_p is usually positive, but its derivative $\frac{\partial \sigma_p}{\partial e_g}$ may be either positive or negative. The second term of equation (76) is positive for the part of the plate current curve which is concave upward and negative for the part which is concave downward. This second term is the most important one which contributes to plate-circuit rectification. k_g and its derivative with respect to e_g are usually positive, so that the first term is usually negative. This is the principal term when a grid leak and stoppage condenser or the equivalent are used. The other terms are usually of less importance than the first two and are in general negative because of the negative signs before them. The relative magnitudes of the several terms will be given in a special

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numerical case. If a gas tube is used the relative magnitudes and signs of the several terms may be very different from those for a hard tube. A more complete discussion of the equations will be given when the experimental results are considered.

When a detector is used in practice some of the impedances included in the general expression (76) may be negligible or absent. At this point, therefore, several special but practical cases of (76) will be examined. In order to make the study more definite the various characteristics of a certain typical triode which contribute to the detection coefficient were experimentally determined and are here presented. The tube used is a Cunningham type 301-A. Figure 7 is a plot of the plate

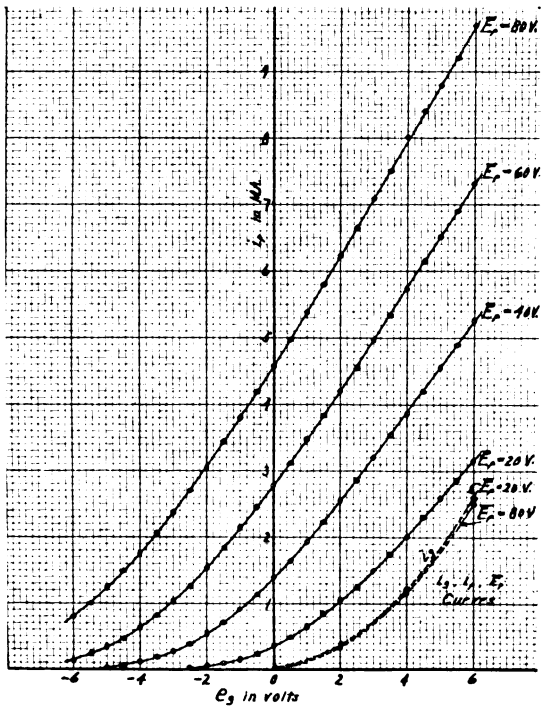

FIGURE 7—Plate Current and Grid Current of Cunningham 391-A Tube

and grid currents. Figures 8, 9, 10, and 11 are, respectively, the values of μ_p, k_p, σ_p, and k_g, of the tube. The values of the derivatives of these curves were obtained graphically. Evidence of a trace of gas is apparent in the irregularities in the course of most of the curves.

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Case 1. Simple Plate Circuit Rectification. To examine the simplest case of rectification we may assume that there are no high-frequency impedances in the plate circuit and no low-

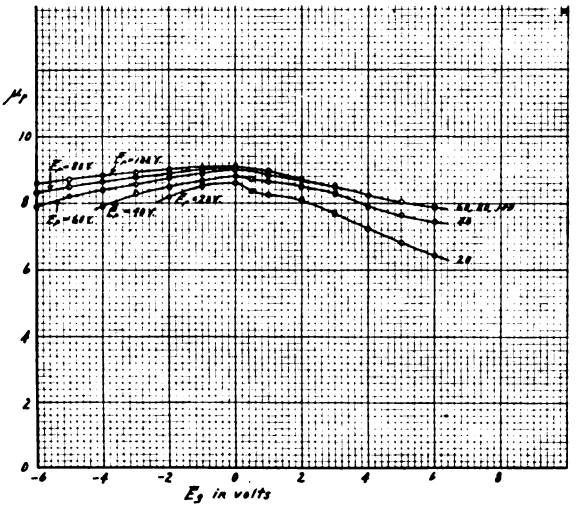

FIGURE 8—Amplification Factor of Cunningham 301-A Tube

frequency impedances in the grid circuit. Figure 12 illustrates this case where it is to be noted that C_b is large enough to be

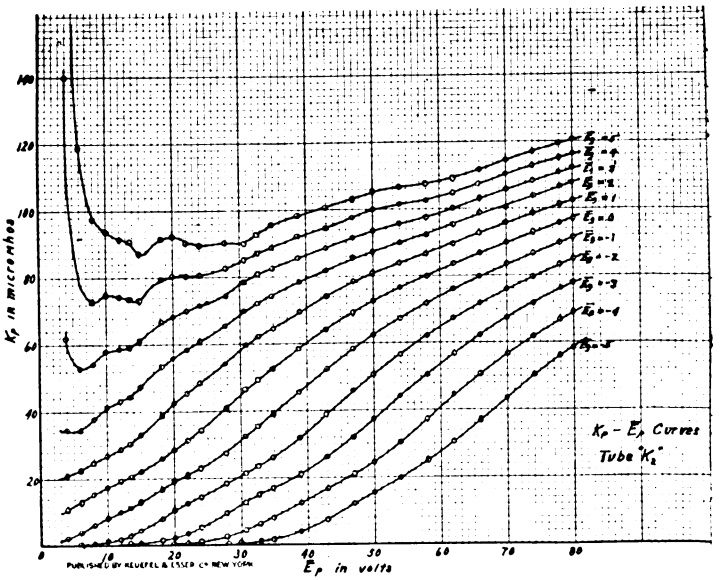

FIGURE 9—Plate Conductance of Cunningham 301-A Tube

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of negligible impedance in comparison with r_p to the high-frequency current. Expressions (72)–(74) now reduce to the following:

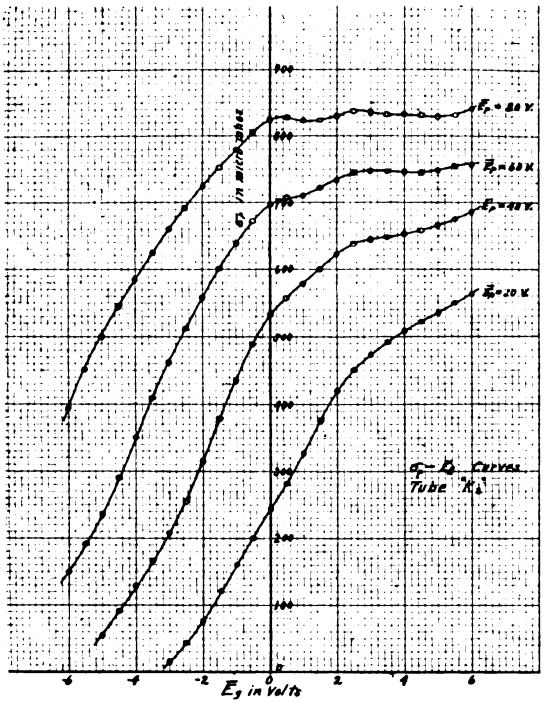

FIGURE 10—Grid-Plate Conductance of Cunningham 301-A Tube

$$\left.\begin{aligned}(\overline{Det.\ I}) &= \frac{r_p}{r_p+\overline{R}_b}\cdot\frac{1}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\\ (\overline{Det.\ E}) &= \frac{r_p}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\end{aligned}\right\} \tag{77}$$

$$\left.\begin{aligned}(\boldsymbol{Det.\ I})_l &= \frac{r_p}{r_p+Z_{bl}}\cdot\frac{1}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\\ (Det.\ E)_l &= \frac{r_p}{4}\cdot\frac{\partial\,\sigma_p}{\partial\,e_g}\end{aligned}\right\} \tag{78}$$

$$\left.\begin{aligned}(\boldsymbol{Det.\ I})_{2l} &= \frac{r_p}{r_p+Z_{b2l}}\cdot\frac{1}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\\ (Det.\ E)_{2l} &= \frac{r_p}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\end{aligned}\right\} \tag{79}$$

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The factor B (see (62)) is practically unity, because R_{ch} and X_{ch} are usually small.

An examination of equations (77)–(79) indicates that in this simple case detection depends upon the value of $\frac{\partial \sigma_p}{\partial e_g}$ or $\frac{\partial^2 i_p}{\partial e_g^2}$. Figure 13 gives the values of $\frac{\partial \sigma_p}{\partial e_g}$ derived from Figure 10. The irregularities in the σ_p curves due to a trace of gas are much more pronounced in the curves of $\frac{\partial \sigma_p}{\partial e_g}$.

Case 2. Plate Circuit Rectification with Tickler Coil. This case differs from Case 1 only in that a radio impedance is

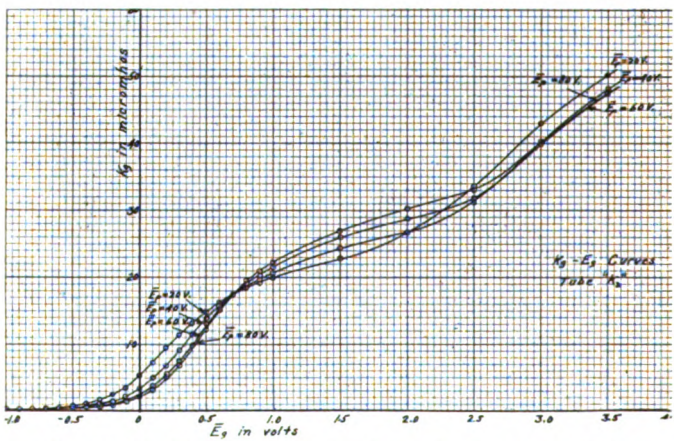

FIGURE 11—Grid Conductance of Cunningham 301-A Tube

added in the plate circuit in the form of a tickler coil. As an approximation, the radio impedance may be considered purely reactive of value $L_p \omega_h$.

The coupling of L_p to the grid circuit reduces the effective resistance of the oscillatory circuit, thereby increasing the value of

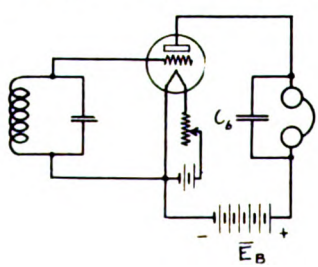

FIGURE 12

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the applied voltage ΔE_g, and in addition, introduces a resistance component into the plate circuit. The value of the resistance added to the plate circuit is usually small in comparison with r_p and is here neglected so that *the coupling does not*

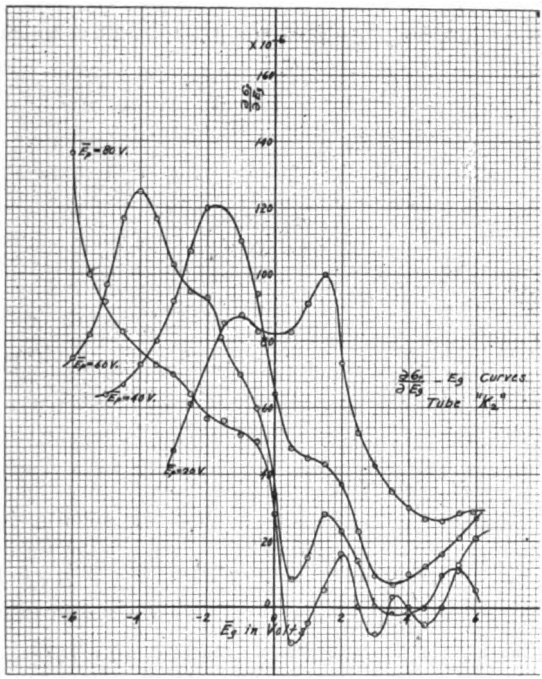

FIGURE 13—$\frac{\partial \sigma_p}{\partial e_g}$ for a Cunningham 301-A Tube

change the value of the detection coefficient but increases the impressed signal voltage.

Equation (76) is in this case the best expression to use and reduces to the following form

$$(\textbf{Det. } I)_l = \frac{r_p}{4(r_p+Z_{bl})}\left[\frac{\partial \sigma_p}{\partial e_g} - 2\frac{L_p\,\omega_h}{\sqrt{r_p^2+L_p^2\,\omega_h^2}}\sigma_p\frac{\partial \mu_p}{\partial e_p} - \frac{L_p\,\omega_h}{\sqrt{r_p^2+L_p^2\,\omega_h^2}}\left(2-\frac{L_p\,\omega_h}{\sqrt{r_p^2+L_p^2\,\omega_h^2}}\right)\mu_p^2\frac{\partial k_p}{\partial e_p}\right] \quad (80)$$

Here again B (see (62)) is assumed to differ little from unity.

We will now examine the relative magnitudes of the three terms of (80) for the particular tube tested. The values of $\frac{\partial \sigma_p}{\partial e_g}$ are shown in Figure 13. The magnitude of $\sigma_p\frac{\partial \mu_p}{\partial e_p}$ plotted

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against grid voltage is given in Figure 14, which shows that the second term is of negligible importance for negative grid voltages. The values of the detection coefficient for positive grid voltages greater than about one are of little interest because, although the magnitude of the detection coefficient may be large for these voltages, the input resistance of the grid circuit is so low that the

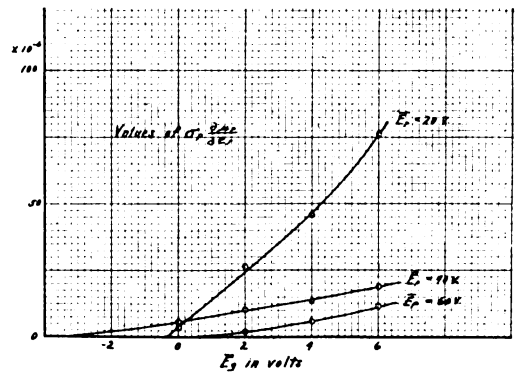

FIGURE 14—$\sigma_p \dfrac{\partial \mu_p}{\partial e_p}$ for a Cunningham 301-A Tube

applied voltage ΔE_o across the oscillatory circuit is much reduced. Figure 15, which gives the values of $\mu_p^2 \dfrac{\partial k_p}{\partial e_p}$, shows that for negative grid potentials $\mu_p^2 \dfrac{\partial k_p}{\partial e_p}$ is of the same order of magnitude as $\dfrac{\partial \sigma_p}{\partial e_g}$, and because of the negative sign before this third term, operates to reduce the detection coefficient. Evidently then a radio-frequency impedance in the plate circuit of a detector

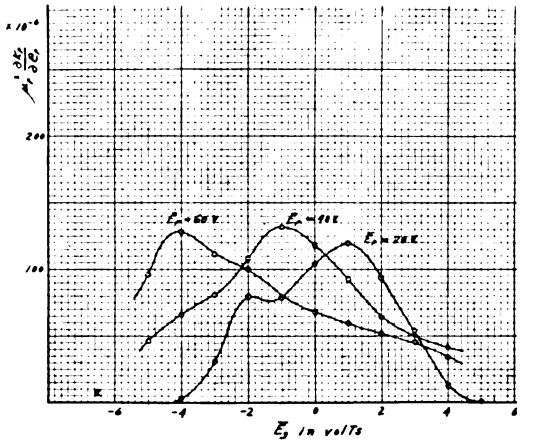

FIGURE 15—$\mu_p^2 \dfrac{\partial k_p}{\partial e_p}$ for a Cunningham 301-A Tube

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operating with plate circuit rectification reduces the effective detection coefficient.

Case 3. Simple Grid-Circuit Rectification. This case, illustrated in Figure 16, applies to the most common use of a triode as a detector. The grid leak R_1 is shunted by the grid

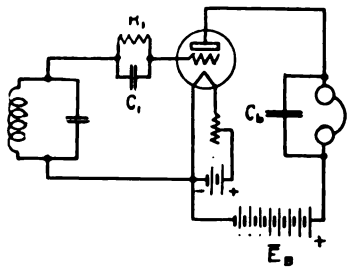

FIGURE 16

condenser C_1. R_1 is usually of the order of from 1 to 10 megohms and C_1 from .00005 to .00025 microfarads. The equivalent series resistance and reactance in the grid circuit are

$$R_c = R_1 \frac{1}{R_1^2 C_1^2 \omega^2 + 1}$$

$$X_c = -\frac{1}{C_1 \omega} \cdot \frac{R_1^2 C_1^2 \omega^2}{R_1^2 C_1^2 \omega^2 + 1} \qquad (81)$$

At radio frequencies the value of R_{ch} is negligible, and $X_{ch} = \frac{-1}{C_1 \omega_h}$, is also usually negligible compared with r_g.

At audio frequencies R_{cl} and X_{cl} have values which vary much with frequency dependent upon the value of $R_1^2 C_1^2 \omega_l^2$ as compared with unity. The table below gives the values of this quantity for different frequencies n when $R_1 = 1$ megohm and $C_1 = .0001$ μf, and also when $R_1 = 2$ megohms and $C_1 = .0002$ μf. The value of $R_1^2 C_1^2 \omega_l^2$ can readily be deduced from the table for other values of R_1 or C_1.

	$R_1^2 C_1^2 \omega_l^2$	
	for $R_1 = 10^6$ ohms	for $R_1 = 2 \times 10^6$ ohms
n	$C_1 = .0001$ μf.	$C_1 = .0002$ μf.
100	.00395	.0631
500	.0988	1.58
1000	.395	6.31
2000	1.58	25.3
4000	6.32	101.
8000	25.3	405.
10000	39.5	631

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Equation (73) reduces in the case now considered to the following:

$$\left.\begin{aligned}(\textbf{Det. } I)_l &= \frac{r_p}{4(r_p+Z_{bl})}\left[-\frac{r_g Z_{cl}}{r_g+Z_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_g}\right]\\ (\textbf{Det. } E)_l &= \frac{r_p}{4}\left[-\frac{r_g Z_{cl}}{r_g+Z_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_p}\right]\end{aligned}\right\} \quad (82)$$

Factor B is again assumed equal to unity.

Evidently when $\frac{\partial \sigma_p}{\partial e_g}$ is positive the second term deducts from the effect of the first term, but since the maximum value of the first term is usually more than ten times greater than the maximum value of the second term, the latter is of little importance. Let us assume that the plate voltage is so adjusted that $\frac{\partial \sigma_p}{\partial e_g}$ is small so that it may be neglected. Then (82) becomes

$$\left.\begin{aligned}(\textbf{Det. } I)_l &= -\frac{r_p}{4(r_p+Z_{bl})}\cdot\frac{r_g Z_{cl}}{r_g+Z_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}\\ (\textbf{Det. } E)_l &= -\frac{r_p r_g Z_{cl}}{r_g+Z_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}\end{aligned}\right\} \quad (83)$$

The magnitudes of $(Det.\ I)_l$ and $(Det.\ E)_l$ in terms of R_1 and C_1 are

$$\left.\begin{aligned}(Det.\ I)_l &= \frac{r_p}{4\sqrt{(r_p+R_{bl})^2+X_{bl}^2}}\cdot\\ &\quad\frac{R_1}{\sqrt{(R_1^2C_1^2\omega_l^2+1)+R_1k_g(2+R_1k_g)}}\sigma_p\frac{\partial k_g}{\partial e_g}\\ (Det.\ E)_l &= \frac{r_p}{4}\frac{R_1}{\sqrt{R_1^2C_1^2\omega_l^2+1+R_1k_g(2+R_1k_g)}}\sigma_p\frac{\partial k_g}{\partial e_g}\end{aligned}\right\} \quad (84)$$

Figure 17 shows the manner in which $\sigma_p\frac{\partial k_g}{\partial e_g}$ varies with $\bar{E}_g$ for a type 301-A tube.

The shape of the detection coefficient curve as plotted against $\bar{E}_g$ is as much dependent upon the fraction

$$F=\frac{R_1}{\sqrt{(R_1^2C_1^2\omega_l^2+1)+R_1k_g(2+R_1k_g)}} \quad (85)$$

as upon the shape of the curves of $\sigma_p\frac{\partial k_g}{\partial e_g}$. Furthermore, the fraction (85) contains the term $R_1^2C_1^2\omega_l^2$, which makes the detection coefficient a function of frequency.

The rapid manner in which F varies with grid polarizing potential for particular plate potentials and values of R_1 and C_1, is

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shown in Figure 18, and the variation of F with frequency is indicated in Figure 19.

The detection coefficient due only to grid-current rectification is equal to

$$(\textbf{\textit{Det. I}})_l = \frac{r_p}{4(r_p + Z_{bl})}\left[F \times \sigma_p \frac{\partial k_g}{\partial e_g}\right] \qquad (87)$$

The value of $(Det.\ I)_l$ for $\overline{E}_p = 60$ volts has been calculated from (87) and is later shown compared with an experimental determination of the detection coefficient for the 301-A tube.

Case 4. Grid-Circuit Rectification with Tickler Coil. The diagram of connections for this case is the same as shown

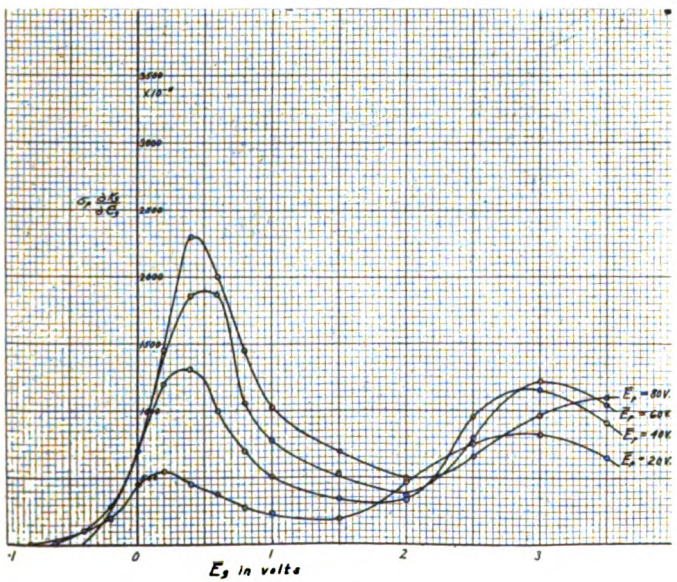

FIGURE 17—$\sigma_p \frac{\partial k_g}{\partial e}$ for a Cunningham 301-A Tube

in Figure 16 except that a tickler coil is added in the plate circuit. The back coupling, as in *Case* 2, in effect decreases the effective resistance of the oscillatory circuit resulting in an increased value of the impressed signal voltage. The tickler coil also introduces a radio reactance in the plate circuit.

The detection coefficient, therefore, depends upon every term of the complete equation (73) or of the better form (76). The first term is much the largest of the four terms. The third term is usually negligible. The second and fourth terms are of the same order of magnitude, but of opposite sign. Therefore, the

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radio load in the plate circuit is in this case an advantage because it increases the value of the fourth term which acts to help out the first or main term.

III—Experimental Determination of Detection Coefficient and Check of Theory

Method of Measurement of Detection Coefficient. The current detection coefficient can easily be measured by a null balance method best explained by reference to the

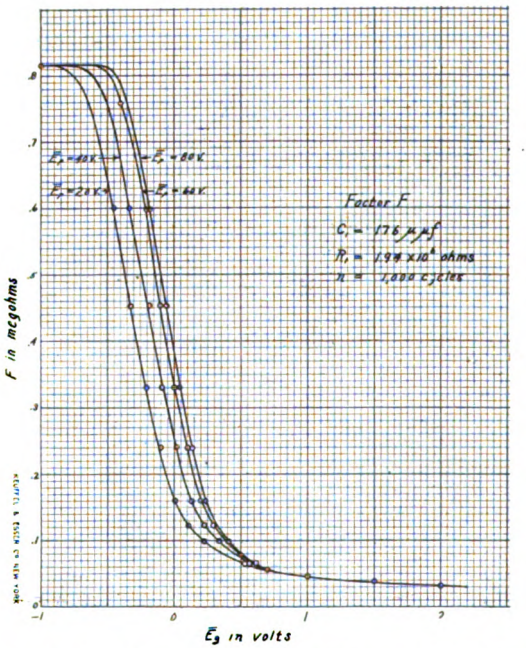

FIGURE 18—Factor F for a Cunningham 301-A Tube

diagram of connections shown in Figure 20. The plate circuit of the detector under test contains the primary winding of an aircore mutual inductance M_b. The primary winding L_b is shunted by a radio by-pass condenser C_b. The grid circuit of the detector tube is connected through the stoppage condenser C_1 and leak R_1, when these are used, and through a potential divider, supplying the grid polarizing potential $\overline{E}_C$, to the potential terminals of a small known four-terminal resistance R_o. The resistance R_o is included in the oscillatory circuit of a radio oscillator so that the oscillatory current multiplied by the resistance R_o gives the radio

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signal impressed on the detector. The plate circuit of the oscillator is fed by a plate battery E_B' in series with a source of

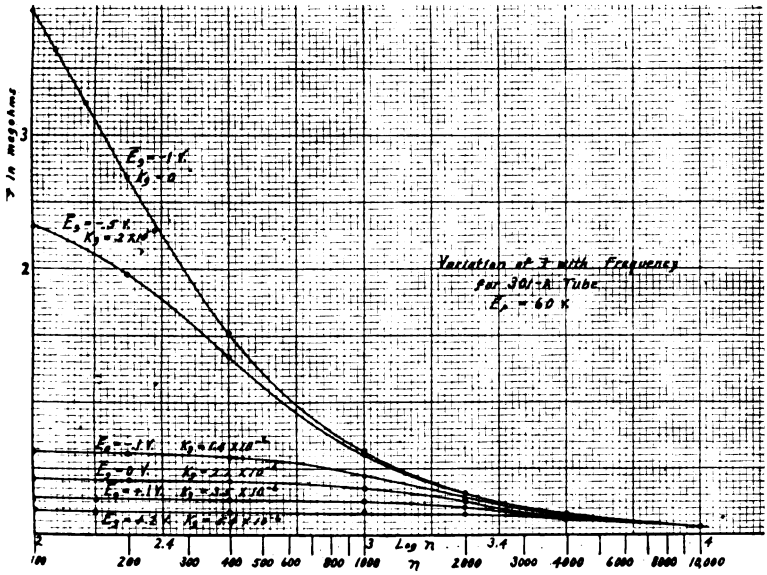

FIGURE 19—Variation of Factor F with Frequency—E_p=60 Volts

1,000-cycle alternating current of known voltage read by voltmeter E_m. A radio by-pass condenser C' is connected across E_m.

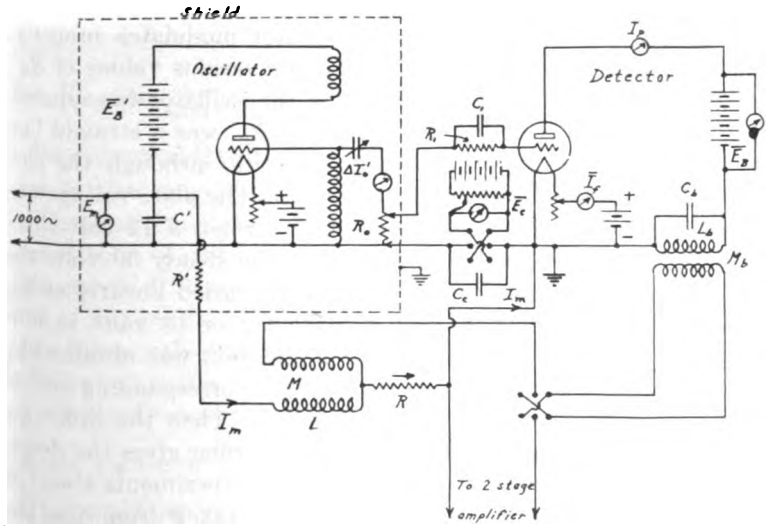

FIGURE 20—Diagram of Connections for Measurement of Detection Coefficient

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This alternating current modulates the radio-frequency current through the resistance R_o. The modulated radio signal acting through the detector gives rise to a demodulated 1,000-cycle current in the primary winding L_b. The induced voltage in the secondary winding of M_b is measured by balancing it in phase and magnitude by the voltage $I_m\sqrt{R^2+M^2\omega_1^2}$ where both R and M are variable. The current I_m is taken from source E_m through a very high resistance R' so that I_m is practically equal to $\frac{E_m}{R'}$. A grounded shield encloses the oscillator as shown in Figure 20.

The values of various quantities as used in the measurements to be described are as follows:

$L_b = .6524\ h$	$R_b = 537$ ohms
$M_b = .6538\ h$	$C_b = .002\ \mu f$
$C_1 = 176\ \mu\mu f$	$R_1 = 1.937$ megohms
$R_o = .914$ ohms	$\lambda = 250$ meters
$E_B' = 30$ volts	$E_m = 12$ volts
$R' = 60{,}240$ ohms	$C' = .010\ \mu f$

M variable up to 10,000 microhenrys (Campbell Mutual Inductance)

R variable in hundredths, tenths, units, and tens.

$C_c = 1\ \mu f$

$\Delta I_o'$ about 30 milliamperes

Oscillator 201-A tube

In order to determine if the oscillator modulates properly, the oscillatory current was measured for various values of $\bar{E}_B'$, while E_m was zero. The back coupling of the oscillator was adjusted until the output current plotted against $\bar{E}_B'$ was a straight line over the range of $\bar{E}_B'$ from 10 to 50 volts, although the plot extended did not pass through the origin. The plate voltage was then set at 30 and it was assumed that when a 12-volt alternating voltage was superimposed upon the steady 30 volts the amplitude of the radio-frequency current varied linearly as the total plate voltage varied from $30-12\sqrt{2}$ or 13 volts to $30+12\sqrt{2}$ or 47 volts. The degree of modulation was obtained by reading from the linear plot the current corresponding to 30 volts and that corresponding to 13 volts. Then the difference between the two currents divided by the former gives the degree of modulation m. This factor was in these experiments about .6. The unmodulated radio current ΔI_o was taken from the plot for $\bar{E}_B'$ equal to 30.

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The value of the audio current in the plate circuit of the detector is obviously

$$\Delta^2 I_p = \frac{E_m \sqrt{R^2 + M^2(2\pi \cdot 1{,}000)^2}}{R' M_b(2\pi \cdot 1{,}000)} \tag{88}$$

Then referring to equation (30), the final expression from which the detection coefficient can be computed is

$$(Det.\ I)_l = \frac{\Delta^2 I_p}{\sqrt{2}\, m R_o^2 (\sqrt{2}\, \Delta I_o)^2}$$

$$= \frac{E_m \sqrt{R^2 + M^2(2\pi \cdot 1{,}000)^2}}{4{,}000 \cdot \pi \sqrt{2}\, m R_o^2 R^1 M_b (\Delta I_o)^2} \tag{89}$$

Since m varies nearly proportionally with E_m, the ratio $\frac{E_m}{m}$ remains practically constant if E_m varies slightly.

The phase of the demodulated plate current can also be determined. If $L\,\omega_l$ is small in comparison with R', then I_m is in phase with the modulation. The mutual inductance M_b introduces a 90° lag so that the angle between the audio plate current and the modulation envelope is

$$\phi = tan^{-1} \frac{R}{M\,\omega l}$$

It is sometimes instructive to plot separately the two components of $(Det.\ I)_l$ obtained from the following relations.

$$(Det.\ I)_{l\,\alpha} = \frac{E_m M}{2\sqrt{2}\, m R_o^2 R'\, M_b (\Delta I_o)^2} \quad \begin{matrix} In\text{-}phase \\ component \end{matrix} \tag{90}$$

$$(Det.\ I)_{l\,\beta} = \frac{E_m R}{4{,}000 \cdot \pi \sqrt{2}\, m R_o^2 R^1 M_b (\Delta I_o)^2} \quad \begin{matrix} quadrature \\ component \end{matrix} \tag{91}$$

Experimental Results. In Figure 21 are plotted the experimentally determined values of current detection coefficient for the Cunningham 301-A tube previously used in the measurement of tube characteristics. These curves are similar in shape to the curves of Figure 13. The solid points shown on the dotted curve of Figure 21 were calculated by equation (78) for $\bar{E}_B = 60$ volts. These calculated points fall on a curve similar in shape to the observed curve but somewhat lower. This discrepancy is not surprising considering the difficulty in deriving graphically accurate values of $\frac{\partial \sigma_p}{\partial e_g}$.

Figure 22 presents the experimental curves of the current detection coefficient for the 301-A tube used with a stoppage

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condenser of 176 $\mu\mu f$. and grid leak of 1.937 megohms. In taking these curves the actual grid potential was determined by short circuiting the grid leak and finding the grid potential which would reproduce the plate current observed when making a determination of detection coefficient. The solid points in Figure 22 were calculated from equation (84) for $\overline{E}_B$ equal to 60 volts.

Figures 23 and 24 show the effect of ionization upon the current detection coefficient. The curves of Figure 23 were taken

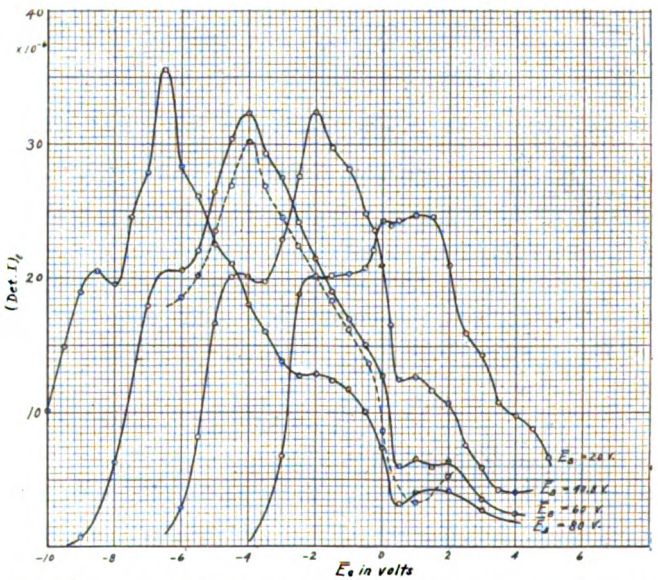

FIGURE 21—Detection Coefficient Curves without Grid Impedance for Cunningham 301-A-Tube

Full lines observed Dotted lines calculated

with a Cunningham Gas Tube type C-300, operating somewhat below normal filament current in order to show temperature saturation of emission. No grid impedance was used so that the detection results from bends and kinks in the plate current curve. The two components of $(\boldsymbol{Det.\ I})_l$ are plotted, $(Det.\ I)_a$ corresponding to the component of audio plate current in phase with the modulating voltage, and $(Det.\ I)_\beta$ corresponding to the component ninety degrees in advance of the modulating voltage. The numerical value of detection coefficient is of course

$$(Det.\ I)_l = \sqrt{(Det.\ I)_a^2 + (Det.\ I)_\beta^2} \qquad (92)$$

Referring to Figure 23, it will be noticed that the plate current curves, particularly the one for 31.5 volts, have peculiar

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shapes. The first rapid rise may be attributed to the neutralization of space change by the positive ions produced by ionization of the gas. The plate current does not continue to rise at the first rapid rate because, due to the small amount of gas present, the supply of positive ions is insufficient to neutralize the rapidly increasing negative space charge and the current flattens out. The second bend, which is concave downward, is due to temperature saturation of emission. The detection coefficient reflects

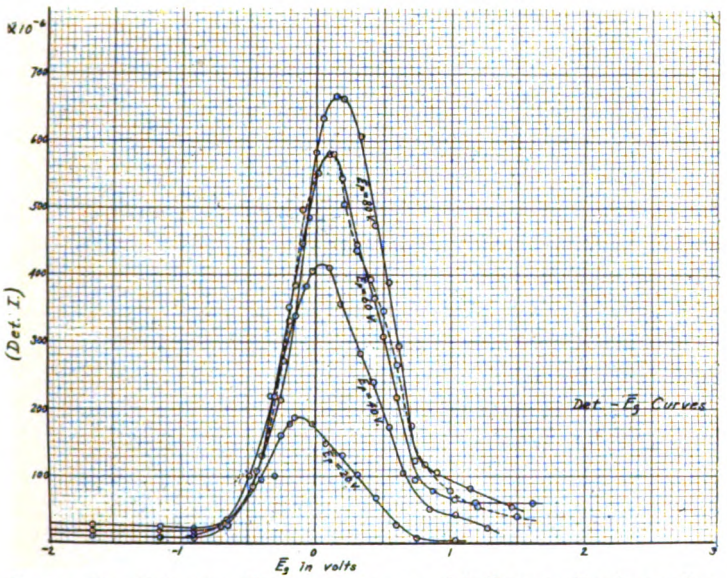

FIGURE 22—Detection Coefficient Curves with Stopage Condenser for Cunningham 301-A Tube

Full lines observed Dotted lines calculated

the several bends in the plate current curves, the $(Det.\ I)_a$ being positive for bends which are concave upward and negative for bends which are concave downward. The very large values of detection coefficient corresponding to the bends which are concave downward can hardly be explained by the curvative of the plate current curves shown. The large values may be due to a time lag in ionization causing the dynamic plate current curve to be different from the static characteristic. The only evidence of kinks in the plate current curves is the small hump on the $(Det.\ I)_a$ curve for 31.5 volts. The curves were taken expressly to show the effect of ionization in changing the rates of curvature of the plate current curves.

The curves of Figure 24 were taken to show the effects of definite ionization kinks in the plate-current curve. The tube

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used contains caesium vapor at low pressure and is known as a UX-200-A. It was operated without grid impedance. One set of curves corresponds to a plate potential of twenty volts, but the other set is for a plate potential of -1.5 volts. The latter curves were taken not because of any practical value in using a negative plate potential, but because of the interest in the fact that electrons do sometimes flow to a negatively charged plate due to contact potentials, and because of the definite ionization kink obtained at these low potentials as evidenced by the

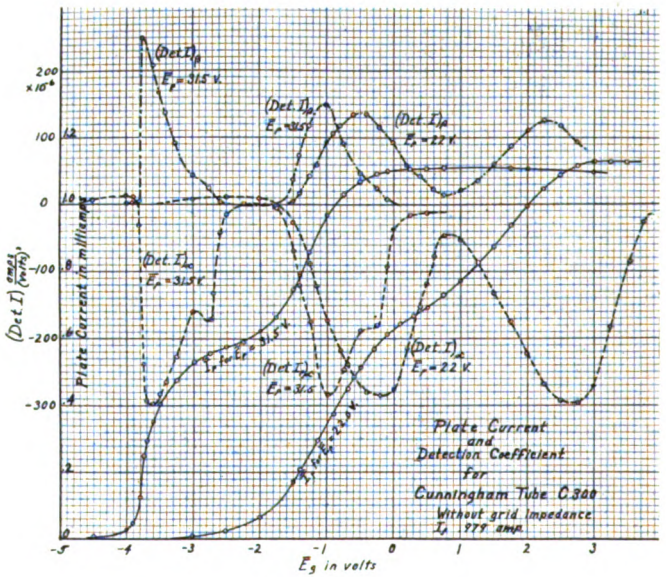

FIGURE 23

shape of the detection coefficient curves. This tube is supposed to be used with a stoppage condenser and grid leak. Under this condition of use, the maximum detection coefficient was found to have the very large value of $7{,}390 \times 10^{-6} \frac{amps}{(volts)^2}$

Although gas tubes have very large values of detection coefficient there is always present more or less hiss, which, to some extent, diminishes their apparent superiority as detectors. Furthermore, their input impedance is usually very low due to their large grid conductance.

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IV—Conclusions

A rectifying detector depends in its action upon a non-linear relation between the instantaneous output current and the instantaneous applied voltage.

If the impressed modulated signal voltage is small, the output contains the following components: a constant current; a current of modulation frequency; a current of double modula-

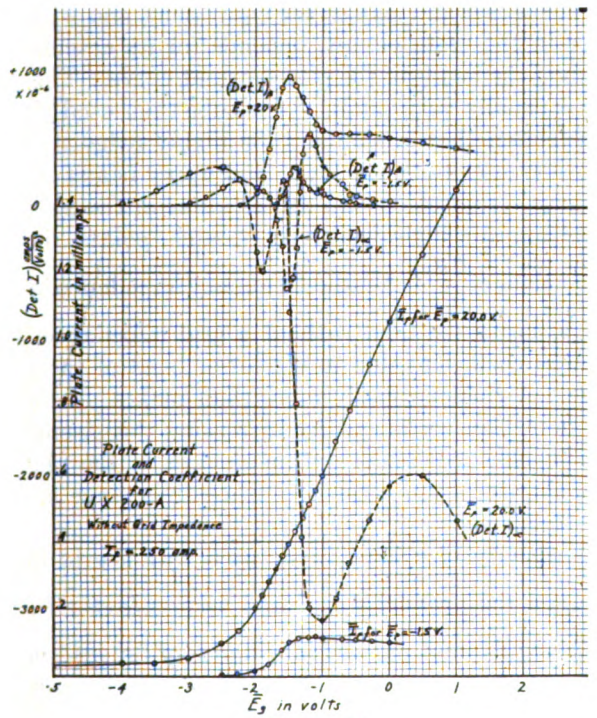

FIGURE 24

tion frequency; and currents of frequencies equal to the sum and difference of the several modulation frequencies. All of these component currents are proportional to the square of the impressed signal voltage. The component of modulation frequency is proportional to the degree of modulation; all others depend upon the square of the modulation or upon the product of the two modulation factors, and hence are small in comparison with the current of modulation frequency, if the degree of modulation is small.

The voltage detection coefficient is a better measure of the detecting action of a rectifying detector because its value is in-

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dependent of the impedance interposed in the plate circuit during the measurement of the detection coefficient. The voltage detection coefficient gives an equivalent voltage which, considered as acting in the plate circuit, gives the audio current which flows through the circuit containing r_p and the plate impedance normally used with the detector. A knowledge of r_p is thus necessary.

A hard triode used without a grid-circuit impedance, that is, stoppage condenser and leak or the equivalent, and with no radio impedance in the plate circuit, depends for its detecting action entirely upon the bends of the plate current curve. The resulting detection is usually very small. A gas tube thus used gives much greater detection than a similar tube highly exhausted because ionization increases both the lower and upper bends of the plate current curve. Ionization also causes kinks in the plate current curve resulting in a high sensitivity at such points due to the very large values of $\frac{\partial \sigma_p}{\partial e_g}$. A radio-frequency impedance in the plate circuit of a hard tube used without a grid impedance decreases the detection coefficient due to the lower bend of the plate-current curve. A tickler usually more than makes up for this decrease in detection coefficient by increasing the strength of the impressed signal. All audio output devices in the plate circuit should be shunted by a condenser whose reactance is small for the radio frequency.

The sensitivity of a hard triode used with a grid impedance depends upon the product $\sigma_p \frac{\partial k_g}{\partial e_g}$ and upon a factor F, which is equal to the equivalent parallel impedance of the grid impedance and the grid-to-filament resistance. The greatest sensitivity obtained for this case when using a hard tube is usually much greater than the maximum sensitivity obtainable without a grid impedance. σ_p should be made large by using the proper plate voltage. The value of $\sigma_p \frac{\partial k_g}{\partial e_g}$ is a maximum for polarizing voltages of a few tenths positive, but F falls so rapidly for positive grid voltages, due to the increase of k_g, that the point of maximum sensitivity is found at a grid voltage more negative than that which gives a maximum value of $\sigma_p \frac{\partial k_g}{\partial e_g}$. The detection coefficient curves are very narrow peaks so it is necessary to adjust the grid polarizing potential to the proper value, usually one or two-tenths of a volt positive. Because of the steady rectified component of current

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in the grid circuit, a strong signal unfortunately alters the polarizing potential.

The ordinary grid leak and stoppage condenser is not the best form of grid impedance because of its variation with frequency and its large value at zero frequency. The ideal impedance is one having negligible resistance to steady currents, a high impedance for frequencies from 100 to 10,000, and low impedance for the radio frequency current.

A tickler coil used with a detector provided with a grid impedance increases the detection coefficient.

The last two terms in the detection-coefficient equation (76) are usually of small importance. At low plate voltages and with gas tubes these terms may, however, contribute largely to the detection-coefficient.

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DISCUSSION ON
LONG DISTANCE RADIO RECEIVING MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1925

(L. W. AUSTIN)

K. Sreenivasan: The curves given by Dr. Austin as a result of his work over a number of years established quite definitely that there are well marked seasonal variations of field strengths of radio signals. The similarity of the curves of monthly averages over a period of four years is certainly striking. But I am a little puzzled to observe the unmistakable gradual increase of field strength of Bordeaux (LY) from year to year as shown in Fig. 5. Can this be a part of a cyclic change extending over a number of years corresponding to any terrestrial or solar phenomenon? Has the author observed similar occurrences with any other stations, say Nauen or Lyons?

The strongest intensity changes of Bordeaux seem to occur anywhere between November and January at Washington and at about the same time at Meudon. That this curve is not an isolated instance is borne out by the observations of R. Mesny and J. Hollingworth. But while the Washington observations on Bordeaux always show an increase in field strength, the curves obtained by Hollingworth for a member of stations show in some cases as marked a reduction in field intensity.

The one point of difference in the two cases is that while the Washington observations represent monthly averages, those of Hollingworth show weekly averages. In some of my observations on Madras (Fort) Radio (295 km. from Bangalore) the following results were obtained for the weekly averages, the working wave length being 4,000 km.

Week ending with	field strength in μv/m
10th October	765
17th "	833
24th "	992
31 st "	611
7th November	664
14th "	779
21st "	982
28th "	788

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I have given these figures to show the marked changes that occur during this period even on 4 km. specially the low values of 611 and 664 $\mu v/m$ for the weeks ending with 31st October and 7th November. While sudden and rapid changes in intensity take place during this period, I do not think, in view of the foregoing, that it will always be an increase in every case.

In this connection, the suggested explanation of Dr. Austin that the agreement of seasonal variations between the observations of Washington and Meudon or Bordeaux is due to causes in the neighborhood of the transmitting station is worthy of note. But in view of Hollingworth's observations and to a little extent of the results on Madras, I am inclined to agree with Hollingworth that this change might be attributed to a rise in the height of the Kennelly-Heaviside layer.

It would have been helpful if Dr. Austin had given us the weekly averages and if and where necessary the daily values of field intensity. The monthly averages have, I think, a tendency to hide the comparatively transient changes which may afford interesting information about the propagation phenomena. But for the derivation of a working formula, the arithmetical work involved in using weekly values would no doubt be very great and in a sense unprofitable.

Quite recently there was reported a severe magnetic storm which had the effect of completely blocking communication from England to Canada by the beam system of short waves. On the other hand, the reception of Annapolis in England showed, it appears, no signs of any departure from the normal. I should like to know if the author observed any abnormality in his readings of the 2nd and 3rd week of October, when the magnetic storm seems to have been strongest. If no abnormality was observed, the explanation for the difference in the case of long and "beam" short waves does not seem to be easy.

L. W. Austin: In commenting on Mr. Sreenivasan's discussion of my paper I should like to point out that there are too many varying factors entering into the changes in signal intensity to permit us to expect to find anything like uniformity from year to year except by averaging over rather long periods.

The causes of the day by day or week by week variations

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may be quite distinct from the prevailing causes of the longer period changes.

The observations made in Washington since 1922 on the European stations at about 3 P. M., E. S. T., when during much of the year part of the signal path lies in darkness, indicate in most cases a slight but regular increase in yearly average intensity, which is most marked in the case of Bordeaux.

These changes, I agree, are probably due to changes in the Kennelly-Heaviside layer either in average height or in ionization, and it seems possible that they may be connected with changes in solar activity. The reality of this relationship can probably only be settled with certainty by regular observations on signal intensity covering at least two sunspot cycles.

In regard to Mr. Sreenivasan's question concerning long wave reception during the second and third weeks of last October, our records do not show any exceptional peculiarities.

Of course we are always glad to send mimeograph copies of our daily signal measurements to anyone interested, as was announced in the "Proceedings," when the regular publication of these measurements was discontinued some years ago.

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DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued November 2, 1926—November 23, 1926

By John B. Brady

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,604,552—HAMMARLUND, L. A., Floral Park, New York. Filed Jan. 10, 1925, issued Oct. 26, 1926. Assigned to The Hammarlund Manufacturing Company, Incorporated.

CONDENSER of variable plate construction where the plates are supported in slotted bearing members which are in turn carried by grooved insulation members constituting the condenser frame.

1,604,986—W. E. GARITY, New York, N. Y. Filed October 7, 1921, issued Nov. 2, 1926. Assigned to De Forest Radio Telegraph & Telephone Co.

ALTERNATING CURRENT GENERATOR in which a construction of tubes is employed having two separated cold electrodes with a restricted electron discharge path therebetween for use in variable light controlling systems.

1,604,987—J. S. GARVIN, New York, N. Y. Filed Dec. 13, 1920, issued Nov. 2, 1926. Assigned to Western Electric Co. Inc.

ENERGIZING CIRCUITS FOR VACUUM TUBES in which direct current generators are employed for supplying the filament and plate circuits of electron tubes and resistances arranged in circuit for controlling the potentials.

1,605,601—F. SCHROTER, Berlin-Schmargendorf, Germany. Filed Mar. 30, 1921, issued Nov. 2, 1926. Assigned to American Telephone & Telegraph Company.

VACUUM VALVE WITH GLOW DISCHARGE where a pair of electrodes are spaced apart by a relatively small unobstructed gap by which a glow discharge may be produced.

1,605,121—F. LOWENSTEIN, New York, N. Y.; John C. Wait, Administrator of said F. Lowenstein, deceased. Filed May 23, 1921, issued Nov. 2, 1926. Assigned to Radio Patents Corp.

RADIO APPARATUS consisting of a wave generator system with simultaneously operative control members for adjusting the several circuits of a signaling system through single actuator.

1,605,230—W. F. HENDRY, New York, N. Y. Filed July 14, 1920, issued Nov. 2, 1926. Assigned to Western Electric Co., Inc.

ELECTRON DISCHARGE DEVICE AND METHOD OF MANUFACTURING THE SAME where a collar is provided for gripping a stem within a tube with blocks of insulating material attached to the collar for supporting the electrodes within the tube.

1,605,333—F. W. DUNMORE, Washington, D. C. Filed Sept. 4, 1925, issued Nov. 2, 1926.

RADIO RECEIVING APPARATUS in which several circuits are tuned simultaneously through a series of independent adjusters operated by a single actuator arranged to move a series of condensers in cooperation with a system of cams.

1,605,607—P. O. PEDERSEN, Frederiksberg, Denmark. Filed June 21, 1919, issued Nov. 2, 1926. Assigned to Poulsen Wireless Corp.

ARC CONVERTOR wherein a pair of arc electrodes are arranged at a distance equal to the most favorable distance of travel of the arc along the cathode.

1,605,608—P. O. PEDERSEN, Frederiksberg, Denmark. Filed June 21, 1919, issued Nov. 2, 1926. Assigned to Poulsen Wireless Corp.

ARC CONVERTOR in which a pair of electrodes are provided for establishing an arc with a plurality of blow pipes for imparting cooling blast to the cathode at a point where the arc is to be extinguished.

Received by the Editor January 26, 1927.

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1,605,627—R. E. THOMPSON, Nyack, N. Y. Filed April 17, 1922, issued Nov. 2, 1926. Assigned One-half to Wireless Improvement Co., and one-half to Wireless Improvement Co., Inc.

RADIO RECEIVING APPARATUS in which a signaling receiving circuit is adjusted over a definite frequency range with an electron tube system having a non-resonant grid circuit with a high resistance winding therein for preventing oscillation while permitting regeneration.

1,605,723—G. W. HEATH, East Orange, N. J. Filed Oct. 11, 1922, issued Nov. 2, 1926. Assigned one-half to Alfred C. Heath.

VARIABLE CONDENSER in which vernier movement for obtaining accurate variation engages with the periphery of one of the rotor plates.

1,605,735—C. W. HOUGH, Boonville, N. Y. Filed Jan. 16, 1926, issued Nov. 2, 1926. Assigned to Wired Radio, Inc.

ELECTRON TUBE SYSTEM wherein the usual house lighting current may be employed for exciting the circuits of the electron tubes, the same tube serving as a rectifier and an amplifier in a signal receiving circuit.

1,605,748—J. L. MARTIN, Amarillo, Texas. Filed Nov. 20, 1925, issued Nov. 2, 1926.

AERIAL AND GROUND CONNECTION FOR RADIO SETS where connections may be made between a radio cabinet through contacts established between the cabinet and connections upon a supporting structure on which the cabinet may rest.

1,605,804—D. F. ASBURY, Washington, D. C. Filed Jan. 18, 1924, issued Nov. 2, 1926.

OPERATING MECHANISM FOR PLURAL TUNING UNIT RADIO APPARATUS where a plurality of condensers may be simultaneously actuated or the condensers may be independently moved through a series of independent adjusters.

1,606,008—H. J. WIEGAND, Milwaukee, Wis. Filed May 29, 1922, issued Nov. 9, 1926. Assigned to The Cutler-Hammer Mfg. Co.

CAPACITANCE CONTROL DEVICE where sets of rollers are arranged for the movement of a flexible conductive sheet from one roller system to another in the variation of electrical capacity.

1,606,184—V. L. Ronci, Brooklyn, N. Y. Filed Dec. 17, 1924, issued Nov. 9, 1926. Assigned to Western Electric Co.

ELECTRON DISCHARGE DEVICE where a collar is provided around a tubular member within an electron tube structure with a bifurcated member straddling the collar and providing an electron support within the tube.

1,606,283—R. WOLF and A. A. WADSWORTH, Jr., of New York, N. Y. Filed Sept. 22, 1922, issued Nov. 9, 1926. Assigned to Hollan S. Duell and Kenneth O'Brien.

RADIO RECEIVING SYSTEM in which an electron tube circuit has its input and output circuits coupled by a free ended inductance connected with the cathode circuit.

1,606,476—J. O. MAUBORGNE and GUY HILL, of Washington, D. C. Filed May 24, 1920, issued Nov. 9, 1926.

RADIO SIGNALING SYSTEM in which a wave coil is arranged for operation as a direction finder.

1,606,660—H. E. MOTT and J. P. MAURER, of Montreal, Canada. Filed Nov. 20, 1925, issued Nov. 9, 1926.

ELECTRICAL CONDENSER of variable plate construction in which the rotor plates are driven through a system of gears for obtaining micrometer adjustment of the condenser.

1,606,773—E. L. NELSON, of East Orange, New Jersey. Filed Aug. 3, 1923, issued Nov. 16, 1926. Assigned to Western Electric Company, Inc.

TRANSMISSION SYSTEM for duplex operation wherein carrier waves of different frequencies are employed and while the frequencies may vary during duplex operation, the difference between the frequencies is maintained constant.

1,606,775—ALEXANDER NYMAN, of Wilkinsburg, Pennsylvania. Filed July 15, 1920, issued Nov. 16, 1926. Assigned to Westinghouse Electric & Manufacturing Company.

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COMBINED RADIO SENDING AND RECEIVING SYSTEM where electron tube circuits are arranged in balanced relationship in such manner that mechanical switching between the transmission and receiving periods is unnecessary.

1,606,791—J. W. HORTON, of Bloomfield, New Jersey. Filed July 18, 1924, issued Nov. 16, 1926. Assigned to Western Electric Company, Inc.

OSCILLATION GENERATOR comprising electron tube circuits arranged for the production of oscillations with means for preventing reaction of the load circuit upon the electron tube circuits comprising a path including resistance, capacity, inductance elements in series and resistance elements in shunt therewith.

1,606,792—F. W. ISLES, of Brooklyn, N. Y. Filed Oct. 6, 1919, issued Nov. 16, 1926. Assigned to Western Electric Company, Inc.

OSCILLATION GENERATOR FOR CURRENT OF CONTINUOUSLY VARYING FREQUENCY consisting of a variable conductor having a rotatable core portion for varying its impedance. The inductance of the inductor is cyclically varied by uniformly and continuously rotating the core portion.

1,606,809—B. Rosenbaum, of Berlin, Germany. Filed Oct. 31, 1921, issued Nov. 16, 1926. Assigned to Westinghouse Electric & Manufacturing Co.

RADIO APPARATUS in which the tuning elements are interlinked by a system of levers for simultaneous operation for controlling the apparatus from a single actuator.

1,606,929—V. A. R. FERNANDES, of Cambridge, Mass. Filed June 6, 1925, issued Nov. 16, 1926.

ARRANGEMENT FOR PRODUCING OSCILLATIONS BY MEANS OF VACUUM TUBES where a vacuum tube path is provided for the current in shunt with the main vacuum tube path for varying the distribution of the current between the two paths at the period of the oscillations.

1,606,940—ROBERT HERZOG, of Berlin, and LEO PUNGS, of Berlin-Charlottenburg, Germany. Filed Aug. 23, 1921, issued Nov. 16, 1926. Assigned to C. Lorenz Aktiengesellschaft.

MEANS FOR KEYING IN RADIO TELEGRAPHY wherein an electron tube circuit is arranged to be controlled for in turn controlling the magnetic condition of an inductance device in the antenna circuit of a transmitter for the formation of signals.

1,606,958—E. PICARD, of Paris, France. Filed June 19, 1925, issued Nov. 16, 1926.

VARIABLE CONDENSER WITH SMALL LOSS wherein silicious insulating material is provided for supporting the frame members of a condenser which in turn carries the plates of the condenser.

1,607,119—O. C. DEUTSCHER, of Brooklyn, N. Y. Filed Aug. 10, 1925, issued Nov. 16, 1926.

CONDENSER of variable construction where a hinged plate of triangular shape may be swung toward or away from a fixed plate.

1,607,158—J. H. HAMMOND, JR., Gloucester, Mass. Original filed Feb. 15, 1917; renewed July 24, 1923; issued Nov. 16, 1926.

MULTIPLEX METHOD AND SYSTEM FOR THE TRANSMISSION OF RADIANT ENERGY in which a series of oscillations are transmitted and impressed upon selectively tuned circuits which operate independently for separating the signaling energy into different channels.

1,607,158—J. H. HAMMOND, Jr., Gloucester, Mass. Original filed Feb. 15, 1918; renewed Oct. 31, 1922; divided Aug. 2, 1923; renewed Mar. 27, 1926; issued Nov. 16, 1926.

ELECTRIC DEVICE in the form of a vapor tube particularly adapted for the generation of high frequency oscillations.

1,607,278—P. C. HEWITT, Ringwood Manor, N. J. Original filed April 23, 1920; renewed July 21, 1926; issued Nov. 16, 1926. Assigned to Cooper Hewitt Electric Co.

SYSTEM FOR THE TRANSMISSION AND RECEPTION OF RADIANT ENERGY for securing secrecy in transmission and reception where a series of wave trains are simultaneously emitted to obtain a predetermined difference in phase relationship between the different waves for selective reception at a correspondingly arranged receiver.

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1,607,277—P. C. HEWITT, Ringwood Manor, N. J. Filed Rec. 1, 1920; issued Nov. 16, 1926. Assigned to Cooper Hewitt Electric Co.

METHOD AND SYSTEM FOR THE TRANSMISSION OF RADIANT ENERGY for secret signaling where a plurality of successive series of periodic impulses are transmitted and received for controlling by their conjoint action, the receiving apparatus.

1,607,277—P. C. HEWITT, Ringwood Manor, N. J. Filed Dec. 1, 1920; issued 1921; renewed July 31, 1926; issued Nov. 16, 1926. Assigned to Cooper Hewitt Electric Co.

SYSTEM FOR ELECTRICAL DISTRIBUTION in which a vapor tube is employed in a circuit for generating high frequency oscillations where the time period of operation of the tube is fixed by a special reactance circuit.

1,607,456—J. H. HAMMOND, JR., Gloucester, Mass. Original filed May 8, 1918; renewed May 1, 1925; issued Nov. 16, 1926.

METHOD OF AND APPARATUS FOR PRODUCING ALTERNATING CURRENTS in which a vapor tube is arranged for the transformation of direct or low frequency current into high frequency oscillations.

1,607,467—M. LATOUR, Paris, France. Filed Dec. 5, 1923, issued Nov. 16, 1926. Assigned to Latour Corporation.

HIGH POWER THERMIONIC VALVE in which the cathode is heated by currents induced therein by the control winding for obtaining electron emission which may be varied for effecting desired control in an associated circuit.

1,607,485—H. SCHMIDT, Berlin, Germany. Filed Feb. 2, 1924, issued Nov. 16, 1926.

WIRELESS SECRET TELEPHONY wherein the frequency of the transmitter is variably controlled and the receiving circuits at distant receiving stations correspondingly controlled in synchronism with the transmitter.

1,607,682—DE LOSS K. MARTIN, West Orange, N. J. Filed Mar. 22, 1924, issued Nov. 23, 1926. Assigned to American Telegraph and Telephone Company.

RADIO REPEATER SYSTEM where the frequency of the received waves is changed at each repeater station and undesirable singing between the repeater stations positively prevented.

1,607,683—DE LOSS K. MARTIN, West Orange, N. J. Filed Mar. 22, 1924, issued Nov. 23, 1926. Assigned to American Telephone and Telegraph Company.

RADIO REPEATER SYSTEM where two way communication may be conducted on different signaling frequencies without interference between the channels.

1,607,837—DE LOSS K. MARTIN, West Orange, N. J. Filed Mar. 22, 1924, issued Nov. 23, 1926. Assigned to American Telegraph and Telephone Company.

RADIO REPEATER SYSTEM in which a plurality of repeater stations are located between terminal stations and communications transmitted therebetween without interference between the channels.

1,607,856—ERNEST E. YAXLEY, Chicago, Ill. Filed Mar. 10, 1926, issued Nov. 23, 1926.

ELECTROSTATIC CONDENSER where a plurality of circular plates are built as a unit in a condenser structure and riverted at the center thereof.

1,607,952—G. HOLST and A. BOUWERS, Eindhoven, Netherlands. Filed Aug. 25, 1925, issued Nov. 23, 1926. Assigned to N. V. Philips' Gloeilampenfabrieken.

ELECTRIC DISCHARGE TUBE where the outer wall of the tube is built up of alternately positioned sections of conductive insulated material.

1,608,003—F. SCHAFFER, Berlin, Germany. Filed June 5, 1924, issued Nov. 23, 1926. Assigned to Gesellschaft fur Drahtlose Telegraphie.

ARRANGEMENT FOR GENERATING AUDIBLE FREQUENCIES IN HIGH FREQUENCY SIGNALING in which an electro-acoustic device is arranged in circuit with a signaling system for controlling the high frequency signaling energy.

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Geographical Location of Members Elected January 5, 1927

Transferred to Fellow Grade

Winfield, L. I., N. Y.,	Pacent, L. G.

Transferred to Member Grade

Bloomfield, N. J., 216 Berkeley Ave.,	Mitchell, C. D.
New York City, 41 Park Row,	Aull, Wilson, Jr.

Elected to Member Grade

Cleveland, Ohio Case School Applied Science,	Martin, J. R.
Melbourne, Australia, 44 Lincoln Road, Essenden,	Bain, J. L.
London, England, General Post Office,	Lee, A. G.
Tonga, Mukualofa,	Land, John R.
U. S. S. R. (Russia), Moscow, Gagarensky Perenlok 2B,	Bashenoff, V. T.

Elected to Associate Grade

El Granada, California,	Kerr, M. F.
Glendale, California, 118 S. Brand Blvd.,	Spencer, B. M.
Los Angeles, California, 757 So. Main Street,	Reed, G. H.
Tujunga, California, P. O. Box 456,	Feringer, W.
Van Muys, California, 14707 Victory Blvd.,	Marsden, G. K.
Hartford, Conn., 46 So. Marshall St.,	Hatry, L. W.
Hartford, Conn., 314 Collins St.,	Page, A. C.
New Haven, Conn., 224 St. Ronan St.,	Weed, C. B.
Stamford, Conn., c-o Electric Specialty Co.,	Berry, E. W.
Stamford, Conn., c-o Electric Specialty Co.,	Haines, W. H.
Wethersfield, Conn., P. O. Box 108, Nott St.,	Mitchell, J. A.
Washington, D. C., Bellevue,	Carboni, A.
Washington, D. C., The Chevy Chase Apts.,	Gross, G. C.
Washington, D. C., Radio Sec., Bureau Standards,	Kirby, S. S.
Washington, D. C., 2633-15th Street, N. W.,	Lutz, J. A.
Washington, D. C., 633 Raleigh Place, S. E.	Merryman, E. A.
Washington, D. C., 2633-15th Street, N. W.,	Taylor, M. B.
Alboin, Illinois, P. O. Box 65,	Woods, M. D.
Chicago, Illinois, 4737 Ingleside Ave.,	Byrnes, W. C.
Chicago, Illinois, 821 Railway Exchange Bldg.,	Crossett, E. C.
Chicago, Illinois, 1076 Roosevelt Road,	Esmaker, J. B.
Chicago, Illinois, 5787 Ridge Ave.,	Gerstle, C.
Chicago, Illinois, 8729 Harvard Ave.,	Kramer, A. W.
Chicago, Illinois, 115 N. Elizabeth St.,	Kruger, L. H.
Chicago, Illinois, 5833 Circle Ave.,	Leverett, G. H.
Chicago, Illinois, 6729 Bosworth Ave.,	Levy, H.
Chicago, Illinois, 5638 S. Washtenaw Ave.,	McCarthy, F. M.
Chicago, Illinois, 4158 Argyle St.,	Moaz, E.
Chicago, Illinois, 3532 W. Adams St.,	O'Byrne, L. C.
Chicago, Illinois, 1647 East 69th St.,	Rose, J. K.
Chicago, Illinois, 2136 Indiana Ave.,	Roscoe, R.
Chicago, Illinois, 230 East Ohio St.,	Slauson, W. T.
Chicago, Illinois, 4734½ Woodlawn Ave.,	Sowell, G. O.
Chicago, Illinois, 7609 East Lake Terrace,	Tyrman, E.
Elmhurst, Illinois, 250 South St.,	Hercher, F. J.
Evanston, Illinois, 804 Ridge Terrace,	Conklin, E. H.
Freeport, Illinois, 130 East Main St.,	O'Connell, G.
Mt. Morris, Illinois, 1½ N. Wesley Ave.,	Tilden, E. R.
Oak Park, Illinois, 815 N. Lombard Ave.,	Downs, K. D.
Urbana, Illinois, 101 S. Busey Ave.,	Chen, C. M.
Clarinda, Iowa, Radio Station KSO,	Maxwell, P. D.
Dubuque, Iowa, 1228 N. Booth St.,	Smith, C. M.
Sabula, Iowa,	Kunua, R. L.
West Oelwein, Iowa, 211-5th Ave.,	Hartley, A. V.
Lawrence, Kansas, Univ. of Kansas,	Kent, P. N.
Wichita, Kansas, 1437 Ellis Ave.,	Andrews, A. E.
Frankfort, Ky., 311 East Main Street,	Culbertson, S. K.
Louisville, Ky., 316 E. Caldwell Ave.,	Marshall, A. B.
New Orleans, La., 217 Custom House Bldg.,	Deiler, T. G.
Baltimore, Md., Room 13, Custom House Bldg.,	Herndon, L. C.
Longmeadow, Mass., 609 Longmeadow St.,	Booth, J. D.
New Bedford, Mass., 162 North St.,	Pierce, C. G.
Roslindale, Mass., 570 Hyde Park Ave.,	Hobart, R. F.
Springfield, Mass., Amer. Bosch Magneto Co.,	Cotter, W. F.
Springfield, Mass., 887 Chestnut St.,	Fox, F. J.
Springfield, Mass., 2787 Main St.,	Laverty, F. W.
Springfield, Mass., 10 Cloran St.,	Vail, H. A.
Wellesley, Mass., Grove Street,	Russell, R. D.
Wollaston, Mass., 55 Franklin Ave.,	Dickman, C. T.
Royal Oak, Mich., 113 W. 4th St.,	Mogredge, J. N.
Detroit, Mich., 3885 Burns Ave.,	Baxter, C. E.
Detroit, Mich.,4464 Cass Ave.,	Robinson, P. S.
Detroit, Mich., 218 Manistique Ave.,	Seymour, F. M.
Detroit, Mich., ss Peter Reiss, Marine P. O.,	Strout, E. M.
Duluth, Minn., 529-4th Ave., East,	England, L. J.
Minneapolis, Minn., 1504 W. Broadway,	Smeby, L. C.
St. Louis, Mo., 4940 Botanical Ave.,	Fowler, D. W.

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Lyndhurst, N. J., 246 Oriental Place,	Gamblin, S.
Orange, N. J., 125 Main St.,	Spencer, R. C.
Westmont, N. J., 112 Ardmore Ave.,	Densham, H. W.
W. Orange, N. J., 2 Harrison Ave.,	Smith, F. L.
Brooklyn, N. Y., 55 Hanson Place,	Bristow, F. R.
Brooklyn, N. Y., 192 Lincoln Place,	Brown, G. G.
Brooklyn, N. Y., 7107-14th Ave.,	Burdick, A. B.
Brooklyn, N. Y., 315 East 96th St.,	Lifschitz, J. L.
Brooklyn, N. Y., 87 Macon St.,	Scofield, R. W.
Buffalo, N. Y., 829 Michigan Ave.,	Brown, S. W.
Buffalo, N. Y., 77 Depew Ave.,	Moore, C. G.
Buffalo, N. Y., 169 Mariner St.,	Morley, W. L.
Jamaica, N. Y., 286 Amherst Ave.,	Steinberger, A. W.
New York, N. Y., 627 W. 43rd St.,	Cockaday, L. M.
New York, N. Y., 1 East 213th St.,	Geise, H.
New York, N. Y., 60 Beaver St.,	Harley, J. B.
New York, N. Y., 133 East 84th St.,	Loscalzo, P. A.
New York, N. Y., 463 West St.,	Mullett, C. B.
New York, N. Y., 120 East 86th St.,	Powley, A. T.
New York, N. Y., 120 Broadway,	Ridgway, J. L.
New York, N. Y., uss Cincinnati,	Riley, C. D.
New York, N. Y., uss Dale,	Wilcox, H. L.
Rochester, N. Y., 1060 University Ave.,	Hanover, E. A.
Rocky Point, N. Y., Radio Corp'n. America,	Barbor, V. H.
Schenectady, N. Y., 177 Nott Terrace,	Roys, H. E.
Schenectady, N. Y., 525 Liberty St.,	Stanton, C. F.
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Lakewood, Ohio, 2225 Olive Ave.,	King, M. V.
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Atglen, Penn.,	Cowan, H. B.
Drexel Hill, Penn., 3232 Marshall Rd.,	Stobbart, A. J.
Johnston, Penn., c-o Star Radio Co.,	Hanson, H. M.
Johnston, Penn., 17 Cox St.,	Reid, A. J.
Philadelphia, Penn., 1533 Pine St.,	Benner, H. S.
Philadelphia, Penn., 1915 S. 12 St.,	Eaton, G. W.
Philadelphia, Penn., 1218 Overington St.,	Eby, J. B.
Philadelphia Penn., Bourse Bldg.,	Hutchinson, S. J.
Philadelphia, Penn., 5442 Sanson St.,	Mahon, F. A.
Philadelphia, Penn., 4422 Chestnut St.,	Wiley, P. F.
York, Penn., 135 W. Jackson St.,	Stewart, J. W.
Aiken, So. Carolina,	Terry, L. M.
Dallas, Texas, 2122 Jackson St.,	Braden, L. O.
Dallas, Texas, 834 No. Marsalis,	Richards, F. O.
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Kent, England, Whitstable, "Hilder Crest,"	Pollard, A. E.
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London, SE19, England, The Vicarage, Gipsy Hill,	Wilson, W. P.
Soest, Holland, Middlewijh St.,	Roorda, G.
Bangalore, India, Indian Inst. of Science, Hebbal P. O.,	Sreenivasan, K.
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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 15 MARCH, 1927 Number 3

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GENERAL INFORMATION

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INSTITUTE ACTIVITIES

NEW YORK MEETING, FEBRUARY 2.

At the February meeting of the Institute, held in the Engineering Societies Building, a paper entitled "Vacuum-Tube Nomenclature" by Dr. E. L. Chaffee was presented. The paper was discussed at length by Dr. Chaffee, J. E. Smith, George R. Metcalfe, J. C. Warner, A. V. Loughren and others. About two hundred members attended the meeting.

FEBRUARY MEETING OF BOARD OF DIRECTION

At the meeting of the Board of Direction, held at Institute Headquarters on February 2, the following were present: Dr. Ralph Bown, President; Dr. A. N. Goldsmith, Secretary; W. F. Hubley, Treasurer; Donald McNicol, Past-President; R. A. Heising, J. V. L. Hogan, R. H. Marriott, L. E. Whittemore and J. M. Clayton, Assistant Secretary.

For the position of Manager, with terms of one year each, the following were appointed by the Board of Direction: J. V. L. Hogan, L. A. Hazeltine and R. H. Manson.

The following Institute officers were appointed by the Board: W. F. Hubley, Treasurer; Dr. A. N. Goldsmith, Editor of Publications and Secretary.

Upon recommendation of the Admissions Committee the following transfers and elections were approved by the Board: Transfer to the grade of Fellow, Frank Conrad. Election to Fellow grade, Dr. K. W. Wagner. Transfer to Member grade: B. W. David, C. E. Drew, Glenn D. Gillett, Virgil M. Graham, Oscar B. Hanson, H. W. Lamson, P. K. McElroy. The following were elected to Member grade: W. E. Barber, C. D. Barbulesco, K. E. Edgeworth, E. H. Freeman, J. D. Graf, L. M. Heron, J. W. Millon, Jr., Miss Mary T. Loomis, E. J. T. Moore and I. P. Rodman.

One hundred and eighteen Associate and eight Junior members were elected.

A geographical distribution of membership, prepared in book form, was presented to the Board by the committee

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composed of E. R. Shute, D. H. Gage, and M. Berger. This list will be useful not only in ascertaining the present membership distribution but also in planning the work of future Sections of the Institute.

Standard forms of procedure to be followed in the establishment of a Section, and a standard Section Petition form, were submitted to the Board by the Sections Committee and were adopted.

Committee Activities

Progress of Committee on Standardization, 1926

Early in 1926 the Institute published a report of the Committee on Standardization which was the result of the work of the Committees of 1923, 1924, and 1925.

During the year 1926 the Committee has carried on its work through five subcommittees, of which the Chairmen are as follows:

1. Vacuum Tubes—L. A. Hazeltine.
2. Circuit Elements—Professor H. M. Turner.
3. Receiving Sets—Dr. J. H. Dellinger.
4. Electro-Acoustic Devices—R. H. Manson.
5. Power Supply—R. H. Langley.

These subcommittees have undertaken the formulation of methods of expressing and measuring the characteristics of radio receiving apparatus and associated circuit elements and devices falling within their respective fields. No general effort is being made to revise the definitions contained in the recently published report, but recommendations will be made as to any new definitions or symbols or any changes which the subcommittees find, during the course of their work, to be desirable.

Coordination of this work with that of other organizations interested in this general subject is facilitated by the presence on this committee of representatives from the Radio Division of the National Electrical Manufacturers' Association, the Radio Manufacturers' Association, and of the Standardization Committee of the American Institute of Electrical Engineers.

The progress made by the several subcommittees during the year 1926 may be summarized briefly as follows:

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1. Vacuum Tubes—L. A. Hazeltine, Chairman.

Meetings of this subcommittee were held in New York on May 28 and July 1, 1926. A draft of a report was prepared based upon the discussions at these meetings. This draft has been submitted to the members of the subcommittee for further consideration and comments are to be submitted in writing to the Chairman. The subcommittee has also under consideration the subject of standard algebraic symbols for the more important vacuum tube quantities.

2. Circuit Elements—Prof. H. M. Turner, Chairman.

The Chairman of this subcommittee has been collecting, by correspondence, the suggestions of the members as to standard methods for measuring the more important characteristics of the elements comprising radio circuits, such as condensers, inductors and resistors. The material will be circulated to the members of the subcommittee for consideration and it is expected that a meeting will be held in the near future.

3. Receiving Sets—Dr. J. H. Dellinger, Chairman.

A meeting of this subcommittee was held in New York on December 1, 1926. A draft has been prepared covering "Tentative Standard Tests for Broadcast Receiving Sets". In addition to the overall tests of "sensitivity", "selectivity" and "fidelity" covered by this draft, the subcommittee has given definite consideration to characteristics of certain portions of receiving sets, and to tests of audio-transformers and of amplifiers. Drafts of proposals covering a number of these questions have been submitted by members of the subcommittee and circulated to others. Consideration of these matters will be continued at the next meeting.

4. Electro-Acoustic Devices—R. H. Manson, Chairman

At the request of the Chairman of the subcommittee, one of the members has prepared a draft of recommendations covering the following aspects of electro-acoustic devices used in radio communication.

a. General Terms and Definitions.

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b. Telephone Receivers.
c. Telephone Transmitters.
d. Performance Indices.

Copies of this draft were sent to the members of the subcommittee for study and comment. A preliminary meeting of the subcommittee was held in New York on January 10, 1927. The Chairman is asking that comments and proposals be submitted to him in writing not later than February 15 so that another meeting of the subcommittee may be held soon after that date.

5. Power Supply—R. H. Langley, Chairman

The Chairman of this subcommittee has been in touch with the work of the corresponding committees of the National Electrical Manufacturers' Association and the Radio Manufacturers' Association and feels that the most satisfactory progress can be made by this subcommittee if it has for its consideration the reports of the mid-winter meetings of these two manufacturers' associations. It is expected that these reports will be available some time in February of 1927. At that time the subcommittee plans to undertake the study of such questions as life testing of batteries used in radio, method of rating batteries for radio purposes, and the specification and measurement of the characteristics of socket-power units.

All of the Standardization work is under the general supervision of L. E. Whittemore, Chairman of the Committee on Standardization, 1927.

Sectional Committee on Radio, A. E. S. C.

The Sectional Committee on Radio, American Engineering Standards Committee, is sponsored by the American Institute of Electrical Engineers and the Institute of Radio Engineers. Its Executive Committee is composed of the following:

Prof. J. H. Morecroft, Chairman; Dr. C. H. Sharp, Vice-Chairman; Dr. A. N. Goldsmith, Secretary and the Chairmen of the Technical Committees.

There are five Technical Committees actively functioning at this time. The chairmen of these are:

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Transmitting and Receiving Sets and Installation, A. F. Van Dyck.

Component Parts and Wiring, L. G. Pacent.

Vacuum Tubes, Dr. J. H. Dellinger.

Electro-Acoustic Devices, L. Espenschied.

Power Supply and Outside Plant, L. W. Chubb.

The report of activities of these Technical Committees since October, 1926 is as follows:

Committee on Transmitting and Receiving Sets and Installation. No formal meetings of the Committee have been held. The following material originating within the Committee is being considered:

1. Methods of classification of radio circuits.
2. Methods of tests of receivers.
3. Use of "preferred numbers".

No material for standardization has as yet been submitted to the Committee by cooperating bodies.

Committee on Component Parts and Wiring. The first general meeting of this Committee was held on December 1, 1926. Those present were: L. G. Pacent, Chairman; Melville Eastham, E. P. Powers, W. F. Hubley and E. L. Hall.

The field of activity of the Committee was outlined by the Chairman. Information was received concerning the various stages to be followed in the formation of a standard, both prior to action by the Committee and subsequently.

The subject of vacuum tubes and sockets being suggested as being proper material for standardization, a subcommittee composed of Messrs. Eastham, Hubley, and Hall was appointed to investigate the work of other Committees to avoid any possible duplication of effort and fields covered by the Committee on Component Parts and Wiring.

Committee on Electro-Acoustic Devices. The first formal meeting of this Committee was held in the office of the Institute of Radio Engineers on November 19, 1926. The following committee members and guests were present: Mr. Whittemore, Acting Chairman and Messrs. Hund, Weyl, Pressley Trainer, Harris, Ringel, Scanlon, Manson, Olney, Frederick, and McKown.

It was agreed that the following devices come within the scope of the Committee:

a. Telephone Receivers.

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b. Loud Speakers.
c. Microphones and Condenser Transmitters.

It was further agreed that the standard symbols adopted by the Institute of Radio Engineers for telephone receiver, loud speaker, and telephone transmitter should be recommended for adoption by the Sectional Committee on Radio, A. E. S. C. The terminal markings adopted by the N. E. M. A. were suggested for adoption by the Sectional Committee on Radio.

The color code to be used in cords was discussed although no decision as to the actual colors was reached.

For the purpose of defining the characteristics of loud speakers the following list was considered for discussion at a future meeting:

1. Transmission frequency characteristics.
2. Efficiency.
3. Frequency range.
4. Volume capacity as limited by:
    a. non-linearity of response.
    b. clearance and mechanical strength of parts.
    c. overheating.
5. D-c. carrying capacity.
6. Insulation breakdown voltage
7. D-c. resistance.
8. Impedance.

It was decided that, in connection with the plotting of transmission frequency characteristics, a logarithmic frequency scale should be used for abscissae.

It is planned to circulate drafts of proposed standards for measuring or specifying the characteristics outlined above, for study and comment.

Committee on Vacuum Tubes. Dr. J. H. Dellinger has been appointed Chairman of the Committee on Vacuum Tubes, succeeding Dr. V. B. Jolliffe, resigned.

No formal meetings of the Committee have been held. It is planned to circulate prints of the exact dimensions and tolerances of the UX-tube base among Committee members as soon as these are available.

Committee on Power Supply and Outside Plant. While no meetings of the Committee on Power Supply and Outside Plant have been held, material for standardization

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from all available services has been collected and distributed to the Committee membership. No action has been taken, however, due to the rapidity of change in design of apparatus having made it difficult to tabulate and work up material for the Committee.

News of the Sections

Los Angeles Section

The Los Angeles Section held a meeting on January 17th in the Commercial Club of Los Angeles. The meeting was addressed by Lee Yount who read a paper entitled "Methods of Explaining Technical Radio to the Layman". The paper was discussed by various members present, including N. E. Brown, D. C. Wallace and W. W. Lindsay, Jr. Fifty members attended the meeting.

Following the discussion of the paper, the annual election of officers was held with the following resutls: Chairman, L. Taufenbach; Vice-Chairman, D. C. Wallace; Secretary-Treasurer, L. Elden Smith; Board of Managers: W. W. Lindsay, Jr., Dr. E. C. Waters and Lee Yount.

The next meeting of the Los Angeles Section will be held at the Commercial Club of Los Angeles on February 2nd. A paper will be read by Dr. MacKeown of the California Institute of Technology, entitled, "Vacuum Tubes".

Connecticut Valley Section

On January 28th a meeting of the Connecticut Valley Section was held in the auditorium of the United Electric Light Company of Springfield, Conn. The speaker was Dr. G. W. Pickard who read a paper "The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism". The paper was discussed by E. A. Laport, Dr. W. G. Cady, Dr. K. S. Van Dyke, G. W. Pettengill and others.

A short talk was given by Alfred Crossley on "Recent Work at the U. S. Naval Research Laboratory".

The attendance at this meeting was thirty-five.

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Philadelphia Section

A meeting of the Philadelphia Section was held on January 28th at the Bartol Laboratories. A paper was read by A. G. Shafer being in the main a discussion of the paper "Vapor Alkali Detector Tubes". General discussion followed. Twenty-six members attended the meeting.

The annual election of officers was held with the following result: Chairman, J. C. Van Horn; Secretary-Treasurer, David P. Gullette. The Chairman appointed a Membership Committee composed of S. J. Hutchinson, A. G. Shafer and H. H. Brown.

Rochester Section

The Rochester Section held a meeting on January 7th at the Sagamore Hotel, Rochester. Francis H. Engel presented a paper on "Vacuum Tubes". The paper was discussed by R. H. Manson and others. The attendance at this meeting was fifty-four.

During the month of February the Rochester Section provided speakers and conducted the noon luncheons of the Rochester Engineering Society. Meetings were held on the first, eighth, fifteenth and twenty-second.

The next meeting of the Rochester Section will be held on March 4, at the Sagamore Hotel. Dr. A. Hund will deliver a paper on "Piezo-Electric Quartz Crystals".

Proposed Section at Detroit

A second preliminary organization meeting looking to the formation of a Section of the Institute at Detroit, Michigan, was held on January 21st in the Federated Radio Trade School. Thomas E. Clark presided. Temporary officers were elected as follows: Chairman, Thomas E. Clark; Vice-Chairman E. D. Glatzel; Secretary, Walter R. Hoffman. The temporary Chairman appointed the following committee Chairmen: Meetings and Papers, A. B. Buchanan; Membership Committee, G. S. Rouston.

The next meeting will be held on February 18th.

Washington Section

A meeting of the Washington Section was held on Feb-

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ruary 9, 1927 in Harvey's Restaurant. The speaker of the evening was Alfred Crossley of the U. S. Naval Research Laboratory. The paper presented was, "Piezo-Electric Crystal Controlled Transmitters". The paper was discussed by Dr. A. Hund and others. The attendance at the meeting was forty.

The next meeting of the Washington Section will be on March 9, 1927 at Harvey's Restaurant. The program will be preceded by a dinner. The speaker will be C. Francis Jenkins who will present a paper on "Visual Radio".

Canadian Section

A meeting of the Canadian Section was held on the evening of February 2, 1927, in the Electrical Building of the University of Toronto, Toronto, Canada. The presiding officer was D. Hepburn. The meeting was addressed by R. H. Coombs, the subject of the paper being "Radio Broadcasting". A general discussion followed in which the effect of harmonic suppressors and the influence on reception of sun spots were brought out.

Twenty-three members of the Section and twenty-nine guests were present at the meeting.

The next meeting of the Canadian Section will be held on March 2, 1927 in the Electrical Building of the University of Toronto. J. M. Thompson will deliver a paper on "Audio Amplification".

Correction

In the paper by A. Crossley on Piezo-Electric Crystal-Controlled Transmitters, in the Proceedings for January, 1927, the development of the crystal oscillator circuit, shown in Fig. 19, was erroneously credited to R. B. Meyer. This circuit was developed by F. B. Monar and was applied by Meyer to the special transmitter system shown in Fig. 20.

Advertising

You have undoubtedly noticed the number of new advertisers who have appeared in each issue of the "Proceedings" since it became a monthly publication. In addition to these new people it has been especially gratifying that

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practically all of our old advertisers have decided to appear regularly in every issue of the new "Proceedings"—that is, twelve times a year instead of six as formerly.

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VACUUM-TUBE NOMENCLATURE

By

E. Leon Chaffee

(Cruft Laboratory, Harvard University, Cambridge, Mass.)

The ease of reading and comprehension of a mathematical discussion is largely dependent upon the simplicity and consistency of the nomenclature. Who has not experienced a certain irritation when reading a mathematical article which is framed in a system of symbols demanding such a useless expenditure of mental effort to remember the meanings of the symbols that little energy is left for the appreciation and understanding of the real truths imparted?

The ideal system of nomenclature is one which demands the least mental effort in distinguishing and remembering the meanings of the symbols. To this end the symbols should be, so far as practicable, suggestive of the quantities that they represent. Initial letters accomplish this aid to memory as, for instance, R is generally used to denote resistance, and C for capacity. Very often, however, the variety of quantities to be represented is so great that the system of using initial letters is inadequate and some other simple and consistent scheme must be used in addition. Some of the symbols of necessity may be arbitrary but whenever possible suggestive symbols should be used. Long usage finally overcomes the objection to an arbitrary symbol but it is surely better to dispense if possible with the period of becoming used to the symbol. I for current is an example of an arbitrary symbol which usage has made a part of the generally accepted nomenclature.

Many branches of science have a standardized system of nomenclature which in some cases is even internationally adopted. The language of expression is then the same in all books and articles. The science of vacuum tubes has unfortunately not yet arrived at this happy state and because of the variety of systems of nomenclature used by different authors it is often difficult to wade through a mathematical article in this subject. Probably the reasons that vacuum tube nomenclature has not become standardized are first,

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because of the comparative newness of the science, and second and most important, because of the peculiar complexity of the situation due to the great variety of quantities which must be distinguished. The Standardization Committee of the Institute of Radio Engineers has fixed upon certain symbols but the author finds the system of nomenclature given by the Committee inadequate and in some cases inconvenient. The author is fully aware of the caution that should be exercised in suggesting new symbols but since the question of nomenclature is still in a state of flux and because the system presented in this paper has received considerable thought and has been found to work well, he believes he is justified in presenting it for discussion at this time.

The general scheme in the system of nomenclature here presented is to represent different kinds of the same quantity as, for instance, steady, instantaneous, average or root-mean-square current by different forms of the same letter with perhaps super or sub symbols to furnish sufficient variety. *Subscripts* are reserved to signify *location*, as I_p means plate current, and also to denote *frequency*, as I_{p1} is a plate current of frequency I. Following out this scheme the meanings of various forms and symbols may now be explained.

Capital letters are used only for quantities which are not directly functions of time, current, or potential but which may be functions of frequency then indicated by the proper subscript.

Small letters represent quantities which are functions of time, or of potential or current. Therefore, instantaneous quantities are always represented by lower-case letters. This has of course been general practice in electrical science.

Bold-faced letters represent vector or complex quantities.

Capital Letter Symbols

Plain capital letters which signify electrical quantities such as current and potential denote the root-mean-square values of sinusoidally varying quantities.

Example: I_p is the r. m. s. value of the sinusoidal part of the plate current. If there are several components of the plate current having different frequencies then it is necessary to specify the frequency by a subscript thus, I_{p1} is the low-frequency component.

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Plain capital letters which signify circuit elements such as resistance, inductance, etc. denote elements which are not functions of time, current or potential but which may be functions of frequency then denoted by the proper subscript.

Example: R, L, C and M represent respectively resistance, inductance, capacity and mutual inductance which are independent of time, current and potential. If a circuit element, say the resistance, depends upon frequency, then R_h is the resistance for the high frequency. The reactance X and impedance Z are of course always functions of frequency which need not be expressed by a subscript unless ambiguity arises.

One dash over a capital letter signifying an electrical quantity denotes an unvarying or steady quantity or component.

Example: $\bar{E}$ is the steady component of the electromotive force represented in Fig. 1.

One dash over a capital letter signifying a circuit element denotes the value of that quantity when a steady current is flowing.

Example: $\bar{R}$ is the value of the resistance offered to a steady current.

Two dashes over a capital letter denotes the time average of the quantity.

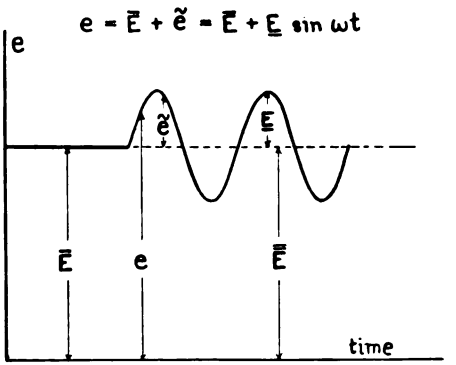

FIGURE 1

Example: $\bar{\bar{E}}$ represents the average value of the non-sinusoidal e.m.f. shown in Fig. 2. In certain cases $\bar{\bar{E}}$ is identical in value with E as in Fig. 1.

One dash below a capital letter signifying an electrical quantity denotes the maximum value of the sinusoidally varying component of the quantity.

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Example: E represents the maximum value of the alternating portion of the e.m.f. shown in Fig. 1.

Two dashes below a capital letter signifying an electrical quantity denotes the peak value of the non-sinusoidally varying quantity.

Example: The significance of $\underline{\underline{E}}$ is shown in Fig 2.

A caret sign over a capital letter denotes the equivalent or effective value of the quantity.

Example: The resistance component of the equivalent impedance of a transformer as seen from the primary circuit is $\hat{R}_1 = R_1 + \frac{M^2\omega^2 R_2}{Z_2^2}$ where subscripts 1 and 2 refer to quantities in the primary and secondary circuits, respectively.

Wavy line or tilde over a capital letter signifying a circuit element denotes the value of that quantity offered to an alternating current.

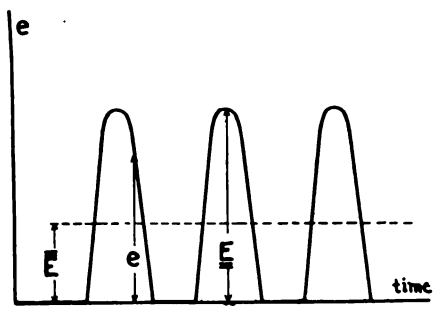

FIGURE 2

Example: $\widetilde{R}$ is the resistance of a circuit to alternating current when it is necessary to signify that it is different from the value offered to a steady current.

Small Letter Symbols

Plain small letters which signify electrical quantities such as current and electromotive force denote total instantaneous values.

Example: In Figs. 1 and 2, e indicates the total instantaneous e. m. f. at any time t.

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Plain small letters signifying circuit elements such as resistance, conductance, amplification factor, etc. which are functions of potential or current are used to denote the values of these characteristics for infinitesimal variations of current or e. m. f. These values thus defined are called *variational* values and each of the quantities is usually the slope of a characteristic curve at some prescribed point.

Examples: If $i=f(e)$ expresses the characteristic curve of a conductor which may be a vacuum tube, then the value of $\frac{de}{di}$ is the variational resistance denoted by r. The reciprocal of r or $\frac{di}{de}$ is the variational conductance denoted by k. The characteristic curve and the value of r are shown in Fig. 3.

The amplification factor μ of a tube is a variational quantity.

The complex impedance of the plate circuit of a triode containing an inductive-resistance load is a function of the plate variational resistance and hence a function of the voltage impressed on the tube. It should, therefore, be de-

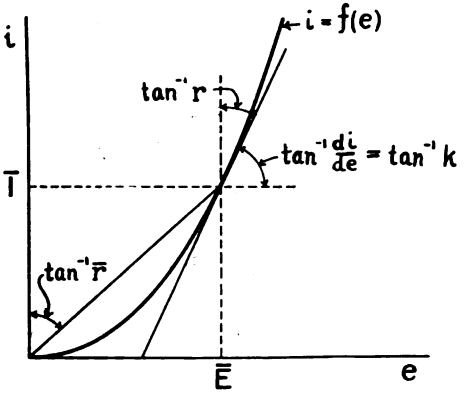

FIGURE 3

noted by a small letter, thus $z=(R+r_p)+jX$ whereas the impedance of the load alone is $Z=R+jX$. The admittance of the load is $Y=G-jB$ but the admittance of the whole circuit is $y=g-jb$, both g and b being functions of r_p.

One dash over a small letter signifying a circuit element, which is itself a function of the impressed e.m.f. or current, denotes the value of the element for steady current.

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Example: If $i=f(e)$ as shown in Fig 3, then $\bar{r}=\frac{\bar{E}}{\bar{I}}$

$\bar{r}$ is obviously a function of $\bar{E}$ but is independent of time.

A wavy line or tilde over a small letter denotes the instantaneous value of a sinusoidal alternating quantity or of the sinusoidal component of a periodic quantity.

Example: Referring to Fig. 1, $\tilde{e}$ is the instantaneous value of the superposed sinusoidal e.m.f. Then $\tilde{e}=\underline{E}\sin \omega t$, and $e=\bar{E}+\tilde{e}$. In Fig. 2, $e-\bar{E}$ is not a sinusoidal e.m.f. but can be developed in a Fourier series, so that

$$e-\bar{E}=\underline{E}_1 \cos \omega t+\underline{E}_2 \cos 2\omega t+$$
$$=\tilde{e}_1+\tilde{e}_2+\tilde{e}_3+$$

Use of Subscripts

Subscripts are in general reserved to denote location and frequency, the frequency subscript always following the location subscript when both are necessary.

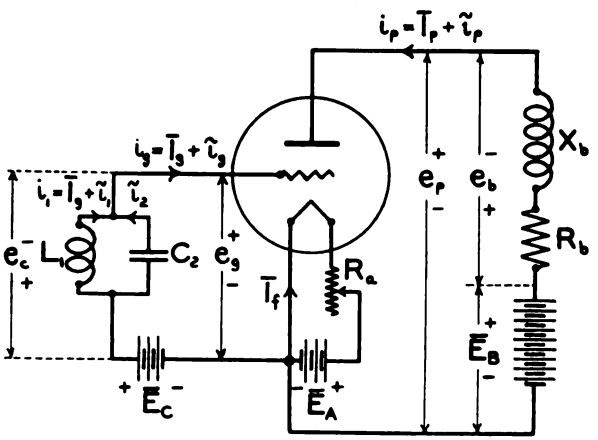

FIGURE 4

Location Subscripts. Particular meanings have been allocated to certain letters used as location subscripts.

Subscripts *f*, *p* and *g* indicate that the quantities to which they are attached are directly associated with the filament, plate and grid, respectively, of a triode.

Example: e_p is the instantaneous plate to filament poten-

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tial. E_f is the steady voltage across the filament. i_g is the instantaneous value of the sinusoidal component of grid current while I_g is the r.m.s. value of this component of grid current and not the r.m.s. value of the total grid current. r_p is the variational resistance of the plate-to-filament path of a triode.

Subscripts *a*, *b*, and *c* are used to denote quantities associated with the series impedances in the filament, plate and grid circuits, respectively, which are external to the triode.

Example: $Z_b = R_b + jX_b$. See Fig. 4. The total instantaneous voltage across Z_b is e_b which is equal to $E_b + \tilde{e}_b$.

Subscripts *A*, *B*, and *C* are used to denote quantities associated with batteries or generators in the filament, plate and grid circuits, respectively.

Example: In Fig. 4, E_B is the plate battery voltage and E_C the grid battery voltage.

Subscripts 1, 2, 3, etc., are used to denote different circuits.

Example: In Fig. 4, the parallel branch in the grid circuit is made up of two parts, one denoted by subscript 1 and the other by subscript 2. Other numbers or letters could equally well be used as subscripts to distinguish the two branches, or prime and second marks might be more convenient in some cases instead of subscripts. From the definition of subscript *c* the symbol Z_c is the equivalent series impedance of the parallel branch. Also

$$\hat{Y}_c = G_c - jB_c = \frac{R_1}{R_1^2 + L_1^2\omega^2} - j\left(\frac{L_1\omega}{R_1^2 + L_1^2\omega^2} - C_2\omega\right),$$ when no ambiguity arises as here the use of the caret sign to denote equivalent value is unnecessary.

Frequency Subscripts. Various subscripts may be used to denote frequency. If, as in modulation and detection, there are two frequencies to distinguish, a high or modulated frequency, and a low or modulation frequency, the subscripts *h* and *l* are convenient. An intermediate frequency can be denoted by subscript *i*. If, on the other hand, harmonics are to be indicated, then the frequencies are all multiples of a fundamental and the numbers 1, 2, 3, etc. can be used to denote the multiple frequencies. In order to differentiate frequency subscripts from location subscripts, it is suggested that the former be represented by heavier type or by italic.

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Example: The plate circuit impedance Z_b may have a different value for the high and low frequencies indicated by Z_{bh} *and* Z_{bl}, respectively. As an example of the use of numbers, the plate current of an oscillator might be expressed as follows:

$$i_p = \bar{\bar{I}}_p + \tilde{i}_{p1} + \tilde{\tilde{i}}_{p2} + \text{- - - - - - - -}$$

Special Symbols

Special letters have become generally accepted as expressing certain quantities. For instance μ is generally accepted as meaning the amplification factor of a tube. Some of these the author includes in the system of nomenclature here presented but others he has rejected as being in his mind unsuited or inconsistent with the system. A list of some of these special symbols is given below with explanatory notes.

The plate and grid currents of a triode are functions of the plate and grid potentials as given by the following expressions

$$i_p = \Phi(\partial_p, \partial_g) \tag{1}$$

and

$$i_g = \Psi(\partial_p, \partial_g) \tag{2}$$

The total differentials of i_p, and i_g are

$$di_p = \frac{\partial i_p}{\partial e_p}\, de_p + \frac{\partial i_p}{\partial e_g}\, de_g \tag{3}$$

and

$$di_g = \frac{\partial i_g}{\partial e_p}\, de_p + \frac{\partial i_g}{\partial e_g}\, de_g \tag{4}$$

The partial derivatives of (3) and (4) have been given significant names and symbols.

$\frac{\partial i_p}{\partial e_p}$ is the *variational plate conductance* and is denoted by k_p. The author objects to the use of g_p for this quantity. In engineering g means the real part of a complex admittance whereas as defined above, k_p is simply the reciprocal of a resistance and k has long been used to denote conductance. It is true that when very short waves are used the time of flight of the electron is appreciable and the characteristics of the tube are then complex quantities. In this case however, equations (1) and (2) are inadequate for time must enter as a third independent variable and then a g_p may properly be defined but not as $\frac{\partial i_p}{\partial e_p}$.

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The reciprocal of $\frac{\partial i_p}{\partial e_p}$ or $\frac{\partial e_p}{\partial i_p}$ is the *variational plate resistance* denoted by r_p.

$\frac{\partial i_p}{\partial e_g}$ *is the variational grid-to-plate conductance* or simply the variational mutual conductance, and is denoted by σ_p. The symbol g_m has been suggested and has to a certain extent been accepted for denoting the mutual conductance. The same objection to the use of g arises here as in the previous case. Furthermore, it has been customary to use g with various subscripts for all of the coefficients in (3) and (4). This is objectionable because the differentiation between mutual conductance and plate conductance, two different quantities although dimensionally the same, depends upon subscripts rather than upon main symbols. A formula containing four kinds of g's differentiated only by subscripts is more bewildering than one in which the differentiation is by main symbols. Furthermore, differentiation of different quantities by subscripts violates the scheme of nomenclature here presented.

Objection might well be raised to the use of the word mutual in a sense which is not in accordance with its strict definition implying equal reciprocal action. There is here no equal reciprocal action but because of long usage it is believed that the meaning of mutual may perhaps be extended to apply to this case.

If the plate current is constant, then $-\frac{\partial e_p}{\partial e_g}$ is denoted by μ_p and is known as the *variational amplification factor*. From relation (3)

$$\mu_p = \frac{\sigma_p}{k_p} = \sigma_p r_p \qquad (5)$$

is the *variational plate-to-grid mutual conductance* denoted by σ_g. $\frac{\partial i_g}{\partial e_p}$ is usually a negative quantity so that with the second negative sign in the definition, σ_g is intrinsically positive. σ_g is called the *variational reflex mutual conductance.*

$\frac{\partial i_g}{\partial e_g}$ is the *variational grid conductance* denoted by k_g. Its reciprocal or $\frac{\partial e_g}{\partial i_g}$ is the *variational grid resistance* or r_g.

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If i_g is held constant then $\frac{\partial e_g}{\partial e_p}$ is the *variational reflex factor* and is denoted by μ_g. According to (4) we have

$$\mu_g = \frac{\sigma_g}{k_g} = \sigma_g r_g \tag{6}$$

We may now write (3) and (4) in the forms

$$di_p = k_p de_p + \sigma_p de_g \tag{7}$$

$$di_g = -\sigma_g de_p + k_g de_g \tag{8}$$

All of the coefficients of (7) and (8) are usually positive quantities.

The inter-electrode capacities of a triode are denoted by C_p, C_g *and* C_m, which are respectively the plate-to-filament capacity, the grid-to-filament capacity, and the mutual or plate-to-grid capacity.

In order to illustrate further the use of the notation, the expressions for the equivalent input admittance of a triode when k_g is negligible are given as developed by the author.

$$\hat{y}_g = \hat{g}_g - j\hat{b}_g \qquad \text{where}$$

$$\hat{g}_g = C_{m\omega} \frac{C_{m\omega} G_b - \sigma_p [B_b - (C_m + C_p)\omega]}{[k_p + G_b]^2 + [B_b - (C_m + C_p)\omega]^2} \tag{9}$$

and

$$\hat{b}_g = -\omega \left\{ C_g + C_m \left[1 + \frac{C_{m\omega}[B_b - C_m + C_p)\omega] + \sigma_p [k_p + G_b]}{[k_p + G_b]^2 + [B_b - (C_m + C_p)\omega]^2} \right] \right\} \tag{10}$$

In (9) and (10) G_b and B_b are the components of the plate load $Y_b = G_b - jB_b$.

The detecting or demodulating quality of a triode has been defined in various ways. The author has developed the theory of detectors in a paper published in I. R. E. Vol. 15, No. 1 and has found the following definitions of detection coefficient convenient.

Just as in the case of amplification, the ability of a tube to amplify depends upon the ampilfication factor μ_p but the actual amplification obtained also depends upon the impedance load in the plate circuit, so in detection certain characteristics of the tube determine the rectifying ability of a tube but the actual rectification obtained depends as well upon the several impedance in the plate and grid circuits and upon a rather complex combination of them with the tube characteristics. There is however some need for an expression involving only the characteristics of the tube

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which can be used as a basis of comparison of different tubes used as detectors. Therefore, two detection coefficients are defined.

In order to make clearer the definitions, the following expressions giving two of the components of the rectified plate current the steady component and the component of modulation frequency, are copied from the article referred to above.

$$\overline{\Delta i_p} = \frac{r_p}{4(r_p+\overline{R_b})}\left[-\frac{\sigma_p r_g \overline{R_c}}{r_g+\overline{R_c}}\cdot\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_g}+\frac{\mu_p^2 R_{bh}^2+X_{bh}^2}{(r_p+R_{bh})^2+X_{bh}^2}\cdot\frac{\partial k_p}{\partial e_p}\right.$$

(11)

$$\left.-\frac{2\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\cdot\frac{\partial \sigma_p}{\partial e_p}\right]\left[\frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(\hat{r}_{gh}+R_{ch})^2+(\hat{x}_{gh}+X_{ch})^2}\right]$$

$$\left(1+\frac{m^2}{2}\right)(\underline{\Delta E_{oh}})^2$$

$$=(\overline{Det.\ 1})\left(1+\frac{m^2}{2}\right)(\underline{\Delta E_{oh}})^2$$

(12)

$$\Delta i_{pl} = \frac{r_p}{4(r_p+Z_{bl})}\left[-\frac{\sigma_p r_g \mathbf{Z}_{cl}}{r_g+\mathbf{Z}_{cl}}\cdot\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_g}+\frac{\mu_p^2(R_{bh}^2+X_{bh}^2}{(r_p+R_{bh})^2+X_{bh}^2}\cdot\frac{\partial k_p}{\partial e_p}\right.$$

(13)

$$\left.-\frac{2\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\cdot\frac{\partial \sigma_p}{\partial e_g}\right]\left[\frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(\hat{r}_{gh}+R_{ch})^2+(\hat{x}_{gh}+X_{ch})^2}\right]$$

$$\sqrt{2m}(\underline{\Delta E_{oh}})^2$$

$$=(\mathbf{Det.\ I})_l\sqrt{2m}(\underline{\Delta E_{oh}})^2$$

(14)

Formulas (11) and (13) are derived for and hence apply *only* for a very small applied radio-frequency voltage in the grid circuit. This is indicated by the Δ sign. The impressed voltage is assumed to be

$$\widetilde{\Delta e}_{oh}=\underline{\Delta E_{oh}}(1+m\sin\omega_l t)\sin\omega_h t$$

where m is the degree of modulation, $\frac{\omega_l}{2\pi}$ is the modulation

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frequency and $\frac{\omega_h}{2\pi}$ is the high or modulated frequency. $(\underline{\Delta E_{oh}})$ is then the maximum value of the unmodulated high-frequency impressed voltage in the grid circuit. $\overline{\Delta^2 I_p}$ of (11) is the change in steady value of plate current represented as of second order magnitude for mathematical consistency.

$\Delta^2 I_{pl}$ is, according to the notation, the r. m. s. value of the sinusoidal component of plate current of modulation frequency $\frac{\omega_l}{2\pi}$; $\overline{R}_c$ is the resistance to steady current of the series load in the grid circuit such as a grid leak resistance, and Z_{cl} is the complex series impedance to modulation frequency of any load in the grid circuit. If a grid leak and condenser are used Z_{cl} is the equivalent series impedance of the parallel combination. R_{bh} and X_{bh} are, respectively, the series resistance and reactance to the radio frequency of any load in the plate circuit. $\hat{r}_{gh}$ and $\hat{x}_{gh}$ are, respectively, the equivalent input resistance and reactance to the high frequency of the triode. R_{ch} and X_{ch} are the series resistance and reactance to the high frequency of any series load in the grid circuit. The other factors have been defined.

In expressions (12) and (14) the current detection coefficients $(\overline{Det.\ I})$ and $(\mathbf{Det.\ I})_l$ are used to denote portions of (11) and (13). These expressions constitute definitions of the new quantities, the *total current detection coefficients.* In words the definitions are given below although the verbal definitions are of questionable value.

The *total steady-current detection coefficient* is the change in plate current per unit squared radio amplitude. $(m=0)$.

The *total current detection coefficient for modulation frequency* is the r. m. s. value of the component of plate current modulation frequency per unit squared radio amplified when the modulation is $\frac{1}{\sqrt{2}}$.

It will be noted that $(\overline{Det.\ I})$ *and* $(\mathbf{Det.\ I})_l$ are identical except for the value of the external grid and plate impedances. Therefore $(\mathbf{Det.\ I})_l$ *becomes* $(\overline{Det.\ I})$ *if* Z_{cl} and Z_{bl} are given their values at zero frequency.

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It is usually preferable to specify the effective fictitious voltage of rectification which if acting in the plate circuit would give rise to the rectified components of plate current given in (11) and (14). These effective voltages are obviously

$$\overline{\Delta^2 \hat{E}} = (r_p + R_{bl})\overline{\Delta^2 I_p} \tag{15}$$

and

$$\Delta^2 \hat{\mathbf{E}}_p = (r_p + \mathbf{Z}_{bl})\ \Delta^2 \mathbf{I}_{pl} \tag{16}$$

Accordingly corresponding total voltage detection coefficients can be defined as

$$(\overline{Det.\ E}) = (r_p + R_b)\ \ (\overline{Det.\ I}) \tag{17}$$

and

$$(\underline{\mathbf{Det.\ E}})_l = (r_p + \mathbf{Z}_{bl})\,(\underline{\mathbf{Det.\ I}})_l \tag{18}$$

Objection might be raised to the cumbersome form of the symbols for detection coefficient. A plain D has been used but this symbol is rejected because of its mathematical use to denote derivative. The form chosen is both suggestive and explanatory, although a trifle awkward. The expression has precedence in the expressions *Div*, *Curl*, etc. and it is used so little in mathematical formulae that this objection of awkwardness is of slight importance.

Returning now to the general expressions (11) and (13), we may assume there is no radio impedance in the plate circuit and no audio or radio series impedance in the grid circuit. Then (11) and (13) reduce to

$$\overline{\Delta^2 I_p} = \frac{r_p}{4(r_p + R_b)} \cdot \frac{\partial \sigma_p}{\partial e_g} \cdot \left(1 + \frac{m^2}{2}\right)(\Delta E_{oh})^2 \tag{19}$$

and

$$\Delta^2 \mathbf{I}_p = \frac{r_p}{4(r_p + \mathbf{Z}_{bl})} \cdot \frac{\partial \sigma_p}{\partial e_g} \cdot \sqrt{2}m(\Delta \mathbf{E}_{oh})^2 \tag{20}$$

These expressions give the rectification due to curvature of the plate current curve. Obviously $\frac{\delta \sigma_p}{\delta e_g}$ is the characteristic of the tube which measures the rectification under these conditions so that we may define a *simple voltage detection coefficient for plate-circuit rectification* for the tube as

$$(det.\ e)_p = \frac{r_p}{4} \cdot \frac{\partial \sigma_p}{\partial e_g} \tag{21}$$

This quantity is independent of frequency and may be conveniently used as a basis of comparing tubes. Note that it is indicated by a small initial letter to distinguish it from the total detection coefficients.

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Tubes are often used as detectors with a grid leak and condenser or the equivalent. If then we assume that there is no plate load for the radio frequency and no grid impedance for the radio frequency in series with the impressed voltage, and also that the rectification due to plate curvature is negligible, then the first term of (11) and (13) is important and they become

$$\overline{\Delta^2 I_p} = -\frac{r_p}{4(r_p+\bar{R}_b)} \cdot \frac{\sigma_p r_g \bar{R}_c}{r_g+\bar{R}_c} \cdot \frac{\partial k_g}{\partial e_g} \left(1+\frac{m^2}{2}\right) (\Delta E_{oh})^2 \quad (22)$$

and

$$\Delta^2 \mathbf{I}_{lp} = -\frac{r_p}{4(r_p+\mathbf{Z}_{bl})} \cdot \frac{\sigma_p r_g \mathbf{Z}_{cl}}{r_g+\mathbf{Z}_{cl}} \cdot \frac{\partial k_g}{\partial e_g} \sqrt{2}m \ (\Delta \mathbf{E}_{oh})^2 \quad (23)$$

Although we can get rid of the $\bar{R}_b$ and $\mathbf{Z}_{bl}$ by expressing the voltage detection coefficient, it is not so easy to make the definition independent of $\bar{R}_c$ and $\mathbf{Z}_{cl}$ except by assuming the series impedance for modulation frequency in the grid circuit infinite. Then we may define the *simple voltage detection coefficient for grid-circuit rectification* as

$$(det.\ e)_g = -\frac{r_p r_g \sigma_p}{4} \cdot \frac{\partial k_g}{\partial e_g}$$

$$= -\frac{\mu_p}{4k_g} \cdot \frac{\partial k_g}{\partial e_g} \quad (24)$$

This expression, too, is independent of frequency and is about the simplest form for use in comparing tubes. Comparative curves of detection coefficient, however, furnish the best basis of comparison and measure of detection property of a triode.

No claim of completeness is made for the system of nomenclature given above. It is presented in the hope that it may serve as a further step toward the early standardization of a system of symbols.

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INFLUENCE ON THE AMPLIFICATION OF A COMMON IMPEDANCE IN THE PLATE CIRCUITS OF AMPLIFIERS*

BY

J. E. ANDERSON

A common impedance in the plate circuits of two or more tubes in an amplifier produces either a decrease or an increase in the amplification. Whether it is the one or the other depends on the type of circuit, on the number of tubes sharing the common impedance, on the phase relations of the various impedances, and on the absolute value of the amplification when the common impedance is zero. For certain values of the common impedance the amplification will increase indefinitely, and the circuit will then oscillate. In general the circuit can only oscillate at one frequency at which the common impedance has the required value.

There are practically no circuits in which there is not a common impedance of appreciable magnitude. Noteworthy examples are: the resistance of a partially exhausted plate battery; the impedance of a plate battery eliminator; the resistance of the balancing potentiometer arms in a-c. heated filament receivers.

If an expression for the amplification in any circuit be obtained, taking into account the impedance in common with two or more plate circuits, it will generally be of the form

$$M=\frac{a+bZ}{A+BZ} \qquad (1)$$

in which A, B, a and b are complex quantities involving the circuit parameters and Z is the common impedance. All of these quantities are functions of the frequency. Whether Z will produce an increase or a decrease in the amplification depends on the signs of the complex quantities associated with Z in (1), particularly on the signs of A and B. Since these two quantities may differ in sign, it is evident that for one value of Z the denominator in (1) will vanish. The

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amplification will then increase indefinitely, and the circuit will oscillate, provided that the numerator does not also vanish at the same time. This it does not do under any practical conditions. Hence the condition for oscillation in the amplifier is that

$$A+BZ=0. \qquad (2)$$

Since this is a complex equation, both the real and the imaginary parts must vanish together. If the common impedance in the plate circuits should satisfy equation (2) at some frequency, the circuit will oscillate at that frequency. In this paper, that value of Z which satisfies this equation will be designated by Z_0 to distinguish it from the actual value of the common impedance. It is independent of the value of Z and is a function of the frequency and of the circuit constants.

If it is only required to determine whether the amplification will be increased or decreased by the common impedance, or whether the circuit might oscillate at some frequency, it is not necessary to obtain the complete expression for amplification, but only the denominator. The effect of the common impedance of any given type may be deduced from the characteristics of Z_0 as obtained from equation (2). Since Z must in all cases be an actual impedance, it must either lie in the first or the fourth quadrant. If it is composed of resistance and inductance it will lie in the first quadrant; if it is composed of resistance and capacity it will lie in the fourth. When Z_0 lies in the first quadrant, the amplification will be increased when the reactance of Z is positive, and the circuit might oscillate at some frequency in the band over which the reactance of Z_0 is positive. Similarly, when Z_0 lies in the fourth quadrant, the amplification will be increased when the reactance of Z is negative, and the circuit might oscillate at some frequency in the band which makes the reactance of Z_0 negative. When Z_0 lies in either the second or the third quadrant, the amplification will be decreased by all possible values of Z, and the circuit cannot oscillate, because then the real part of Z_0 is negative, which would require a resistance of the same sign in the common impedance in order to satisfy equation (2). In some circuits Z_0 will never enter either the first or the fourth quadrant; in others it will never enter the second or the third; and in still others it may move throughout the four quadrants as the frequency varies.

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Analysis of Two-Tube, Transformer-Coupled Amplifier

The two-tube, transformer-coupled amplifier shown in Fig. 1 will be first considered. Let μ_1 be the amplification constant of the first tube, E_1 the input voltage to that tube, i_1 the a-c. component of the plate current in the tube, z_1 the

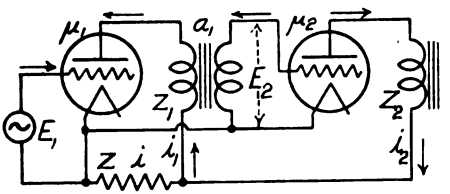

Figure 1

total impedance in the plate circuit exclusive of the common impedance, Z_1 the load impedance, also exclusive of Z. Let the corresponding quantities in the second stage be subscripted by 2, but otherwise the same. Let the current in Z be denoted by i. The effective emf in the first plate circuit is $-\mu_1 E_1$ and that in the second, $-\mu_2 E_2$, the negative signs being employed to account for the fact that the tubes change the phase of the input voltage. The input voltage E_2 to the second tube is equal to $aZ_1 i_1$, where a is the transformer ratio, assumed to be constant and positive. The current in the common impedance is the sum of the two plate currents. Hence the following equations result:

$$\left.\begin{aligned} -\mu_1 E_1 &= z_2 i_2 + Z i \\ -\mu_2 E_2 &= z_1 i_1 + Z i \\ E_2 &= a Z_1 i_1 \\ i &= i_1 + i_2 \end{aligned}\right\} \qquad (3)$$

These reduce to

$$\left.\begin{aligned} -\mu_1 E_1 &= (z_1 + Z) i_1 + Z i_2 \\ 0 &= (a\mu_2 Z_1 + Z) i_1 + (z_2 + Z) i_2 \end{aligned}\right\} \qquad (4)$$

The amplification in the circuit may be defined as the ratio of the voltage across the final load to the initial input voltage. In this case, then, $M = Z_2 i_2 / E_1$. Since only the final current is involved, it is only necessary to solve equations (4) for i_2 in order to get the amplification. The solution is

$$M = \frac{\mu_1 Z_2 (a\mu_2 Z_1 + Z)}{z_1 z_2 + Z(z_1 + z_2 - a\mu_2 Z_1)} \qquad (5)$$

When the denominator in (5) is zero the amplification is infinite, and therefore the condition for oscillation is

$$Z_0 = \frac{z_1 z_2}{a\mu_2 Z_1 - (z_1 + z_2)} \qquad (6)$$

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From both equations (5) and (6) it is apparent that the common impedance causes an increase in the amplification and that oscillation at some frequency is possible, provided that the resistance in the common impedance is high enough.

In defining voltage E_2 it was assumed that the transformer ratio was positive, or such that the phase of the primary and secondary voltages remains the same. When a pair of leads to the transformer is reversed, that is when the sign of a in equation (6) is changed, all the terms in the denominator of (6) become negative. This also makes Z_o negative. Then the circuit can no longer oscillate at any frequency, and there will be a decrease in the amplification. This is exactly the reverse from the case of intentional inductive feed-back, which requires a negative ratio for oscillation. The reason for the difference is not difficult to find. In Fig. 1 the in-phase components of the currents and voltages have been indicated by arrows in the case of positive ratio. It will be seen that part of i_2 is forced by Z to back up through the primary of the transformer and through the first tube, and that this part of i_2 is in phase with i_1. The primary current and voltage are thus increased, and consequently the amplification is also increased. When the ratio is negative all the arrows pertaining to the second tube are reversed. Currents i_1 and i_2 will be in phase in Z. This will increase the voltage drop in Z, and consequently the current circulating in the plate circuit of the first tube will be decreased. This in turn will decrease the primary voltage and the amplification. In the case of ordinary inductive feed-back i_2, or a part of it, comes through the primary of the transformer in the opposite direction, and thus the ratio must be negative if i_1 and i_2 are to be in phase.

Since all the impedances entering into Z_o in equation (6) are inductive, one might assume that the reactance of Z_o is positive, but that is not necessarily the case. For some frequencies it may be negative. The angle of Z_o, or Ψ_o, is equal to the sum of the angles of the two factors in the numerator diminished by the angle of the denominator. The negative term in the denominator has a negligible effect on the angle of the denominator, and therefore this angle is that of Z_1. If the angles of z_1 and z_2 be θ_1 and θ_2, respectively, and that Z_1 be Φ_1, then $\Psi_o = \theta_1 + \theta_2 - \Phi_1$. For low frequencies the negative term will exceed the two posi-

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tive terms. Oscillation is then possible, and the amplification will be increased, when the common impedance consists, for example, of a resistance with a condenser across it. For frequencies above that which makes $\Psi_0=0$, oscillation is possible, and the amplification will be increased, when the common impedance consists of resistance and inductance. When $\Psi_0=0$, pure resistance only is required for oscillation, and at the frequency where that occurs oscillation is most probable because the absolute value of Z_0 is then a minimum.

In a circuit having no more than two reactive branches it is easy to calculate the frequencies at which Ψ_0 will be—90 deg., 0 deg. and 90 deg., that is, to predetermine the frequency bands in which the amplification will be increased when the common impedance has negative or positive reactance. It may thus be proved from equation (6) that Z_0 can never lie in either the second or the third quadrant. Also it may be proved that if the sign of the transformer be made negative there is no frequency which will bring Z_0 into the first or the fourth quadrant.

It is interesting to examine equation (6) graphically.

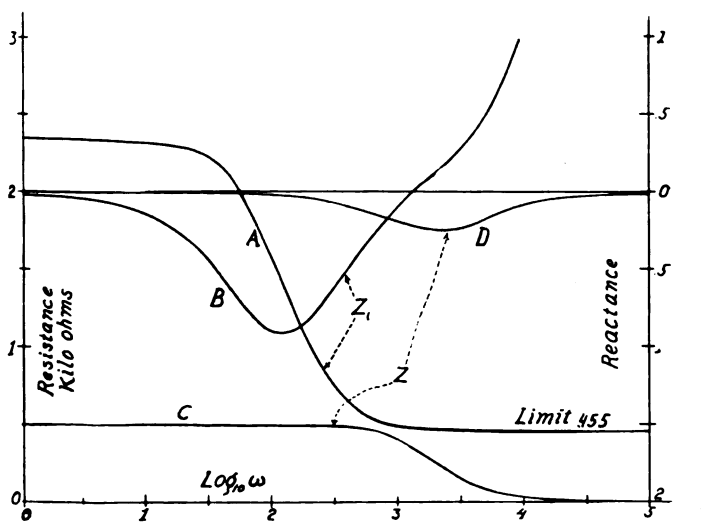

Figure 2

In Fig. 2, Curve *A* is the locus of the real part of Z_0 and Curve *B* is the locus of the imaginary component, plotted as functions of the common logarithm of the angular velocity of the current vector. In computing the data for these

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curves the following values of impedances and constants were used: $z_1=20{,}000+j25\omega$, $z_2=18{,}000+j5\omega$, $Z_1=4{,}000+j25\omega$, $a=6$, and $\mu_2=8$. The scale at the left refers to the real part and that at the right to the imaginary components.

The locus of the real part of Z_0 starts with a high value for low frequencies, drops rapidly between 31.6 and 1,000 radians, and finally approaches the limiting value 455 ohms as the frequency increases. The locus of the reactance starts at zero for zero frequency, decreases to a minimum just above 100 radians, and then rises as the frequency increases. It crosses the axis of zero reactance at about 1,300 radians. It is the region in which the reactance is negative that is of chief interest, because in most receivers the common impedance consists of the resistance of the "B" battery with some intentional or distributed capacity across it, the angle of which is always negative.

The loci of the real and imaginary parts of an impedance, Z, consisting of 500 ohms shunted with a one microfarad condenser are also plotted in Fig. 2. Curve C is that of the real and Curve D that of the imaginary. Below 316 radians the effective resistance is practically 500 ohms; above 10,000 radians it is very small. The reactance of the impedance is very small for both high and low frequencies, but it reaches a negative maximum at the frequency where the rate of change of the real part is greatest.

At a point, just below 1000 radians, Curves B and D intersect. At this frequency, then, the first condition for oscillation is satisfied, because the reactance of the common impedance is equal to the reactance of Z_0. The circuit will oscillate at this frequency provided the second condition is also satisfied. If it does not oscillate at the point, there will be a prominent peak in the amplification, the height of which depends on how nearly the second condition is satisfied.

Curves A and C do not intersect at any point, and therefore the second condition for oscillation is not satisfied. However, they approach each other very closely at one point, and the frequency where this occurs is very close to the frequency where the two reactances are equal. The amplification peak will therefore be high. There are three ways in which to make Curves A and C intersect, and hence to cause oscillation in the circuit. In the first place the resistance in the common impedance may be increased. This will

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raise Curve C for all frequencies. For one value of the resistance the two curves will be tangent at one point; for higher values of the resistance there will be two points of intersection. Intersection of the two curves may also be caused by reducing the capacity of the condenser across the resistance, provided, of course, that the resistance used is greater than the limiting value of the real part of Z_o. With the smaller capacity across the resistance Curve C will remain close to the 500 line for higher frequencies. Intersection may also be brought about by increasing the amplification, that is, by increasing μ_2 and a. This will lower the right hand portion of Curve A. If the amplification is high enough there will be an intersection no matter how close to the axis Curve C may be.

Two-Tube, Direct-Coupled Amplifier

All the equations used in the analysis of Fig. 1 apply in the case of a two-tube, direct-coupled circuit such as is shown in Fig. 3, with the exception of the definition of the input voltage to the second tube. In this case $E_2=Z_1i_1+Zi$,

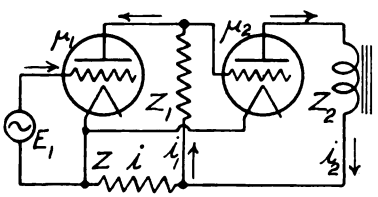

Figure 3

which is the sum of the voltage drops in the coupling and common impedances. If this change be made in equations (3) they reduce to (7), in which $Y_2=(\mu_2+1)$

$$-\mu_1E_1=(z_1+Z)i_1+Zi_2$$
$$0=(\mu_2Z_1+Y_2Z)i_1+(z_2+Y_2Z)i_2 \qquad (7)$$

Equations (7) yield the following expression for amplification and for the critical value of Z:

$$M=\frac{\mu_1Z_2(\mu_2Z_1+Y_2Z)}{z_1z_2+Z(R_1\mu_2+z_1+z_2)} \qquad (8)$$

$$Z_o=\frac{-z_1z_2}{(R_1\mu_2+z_1+z_2)} \qquad (9)$$

If the circuit in Fig. 3 is resistance-coupled there is only

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one reactance branch, that containing z_2. The angle of the denominator of (9) is very small no matter what the angle of z_2 or θ_2, may be. Hence $\Psi_0=180°+\theta_2$, that is, Z_0 lies in the third quadrant when the reactance of z_2 is positive and in the second quadrant when it is negative. Hence when the circuit in Fig. 3 is resistance coupled it cannot oscillate at any frequency, and the amplification will be decreased for all frequencies. Why this is so is obvious from Fig. 3. Current i_2 flows through Z in the opposite direction from i_1, and hence it reduces the input voltage to the second tube, since $Zi+Z_1i_1$ will be less than Z_1i_1 alone.

When the coupling in this circuit is a choke coil instead of a resistance, there may be a frequency band in which Z_0 will lie in the fourth quadrant. This occurs when $\theta_1+\theta_2$ is considerably greater than 90 deg., and before the angle of the denominator has become large. The larger $R_1\mu_2$ is the wider will be the band in which oscillation might occur. However, since both z_1 and z_2 will have considerable distributed capacity it is probable that in a practical case the angles will not become large enough to project Z_0 into the fourth quadrant, and then the circuit would be stable for all frequencies.

Analysis of Multitube Amplifiers

In the system of simultaneous equations applying to any given circuit there must be as many currents and as many voltage equations as there are tubes sharing the common impedance. All the voltage equations will be of the form $-\mu_nE_n=z_ni_n+Zi$, and the current summation will be of the form $i=\Sigma i_m$. The auxiliary voltage input equations will be in the form of either $E_{n+1}=a_nZ_ni_n$ or $E_{r+1}=Z_ni_n+Zi$, depending on whether the coupling between the nth and the $(n+1)$st tube is by transformer or by impedance. If the circuit has more than two tubes, both types of coupling may occur in the same amplifier.

To obtain the complete solution for the amplification in a circuit having more than two tubes entails a great deal of work. However, if only the condition for oscillation in the circuit is required, the work may be shortened, because this condition is simply that the determinant of the system of equations shall vanish. It is only necessary to arrange the coefficients in determinant form, write out its value,

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and equate it to zero. In applying the method it is quickest to regard the current in the common impedance as one of the unknown and make the determinant of one order higher than the number of tubes involved.

THREE-TUBE, TRANSFORMER-COUPLED CIRCUIT

The determinant for a three-tube, transformer-coupled circuit takes the following form:

$$\begin{array}{r|cccc|} & i_1 & i_2 & i_3 & i \\ -\mu_1 E_1= & z_1 & o & o & Z \\ o= & a_1 Z_1 \mu_2 & z_2 & o & Z \\ o= & o & a_2 Z_2 \mu_3 & z_3 & Z \\ o= & -1 & -1 & -1 & 1 \end{array} \qquad (10)$$

Currents are written at the top of the columns and the voltages at the left of the rows simply to identify the coefficients. The last row is the current summation equation. This determinant may be reduced by making use of the property that any line, row or column, may be added to, or subtracted from, any other parallel line without changing the value of the determinant. The value of the above determinant will be found to be

$$D=z_1 z_2 z_3+Z[z_1 z_2+z_2 z_3+(a_1 Z_1 \mu_2-z_1)(a_2 Z_2 \mu_3-z_3)] \qquad (10b)$$

whence

$$Z_o=\frac{-z_1 z\ z_3}{z_1 z_2+z_2 z_3+(a_1 Z_1 \mu_2-z_1)(a_2 Z_2 \mu_3-z_3)} \qquad (11)$$

is obtained as the condition for oscillation by placing D equal to zero.

No definite conclusions as to the probability of oscillation in this three-tube, transformer-coupled circuit can be reached without substituting actual values of constants and impedances. However, the apparent sign of the denominator in (11) is always positive and the apparent sign of the numerator always negative. Hence it would seem that this circuit will not oscillate as readily as the two-tube circuit when the ratio in that circuit was positive. That this might be so is borne out by the directions of the currents in Fig. 4a. Currents i_1 and i_3 flow in the same direction

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while current i_2 flows in the opposite. Hence i_2 aids i_1, and i_3 aids i_2 but opposes i_1. Due to the relative values of the currents, the opposition just about neutralizes the conjunction. But on account of phase relations introduced by the impedances there may be a certain frequency band over

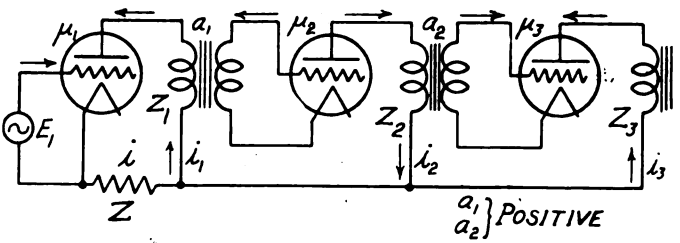

Figure 4a

which the amplification is increased by the common impedance, as actual substitution has shown. It is not safe to trust to the apparent sign of Z_0.

When both of the transformer ratios are negative, the signs of all the terms in the denominator of (11) are positive, while the sign of the numerator remains negative. Then Z_0 apparently lies in either the second or the third

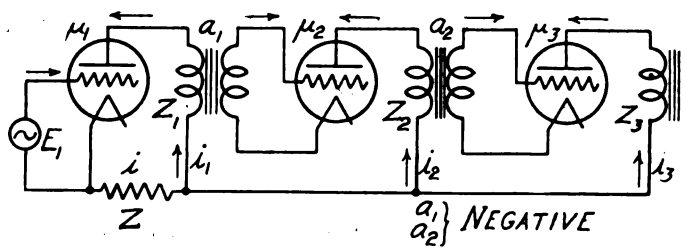

Figure 4b

quadrant, and one would expect a stable circuit. In Fig. 4b the in-phase components of the currents have been indicated. All flow in the same direction, and therefore both i_2 and i_3 tend to reduce the amplification. If there is any band in which the amplification will be increased it can only be because the phase relations of the impedances are such as to bring Z_0 out of the second or third quadrant into the quadrant in which Z lies.

If the two transformer ratios have different signs, the apparent sign of Z_0 becomes positive. The two cases show

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about the same propensity to oscillate, both as judged from the formula and from the circuit diagrams Figs. 4c and 4d, and both are apparently less stable than either of cases 4a or 4b.

Four-Tube, Transformer-Coupled Circuit

The determinant for the solution of a four-tube, transformer-coupled amplifier may be obtained from (10) by adding another row and another column to that determineral q is defined $q_n = a_n Z_n \mu (n+1)$.

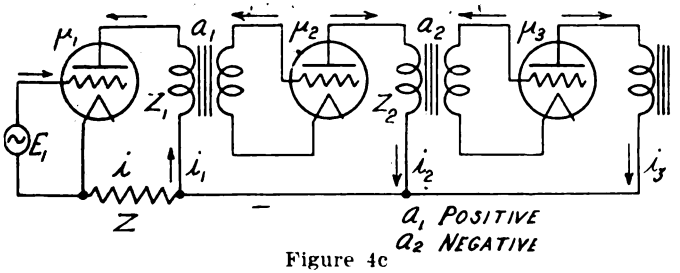

Figure 4c

and. The row to be added is $0 = a_3 Z_{3\ 4} i_3 + z_4 i_4 + Zi$ and the column is $0,0,0,z_4,-1$. This fifth order determinant yields the following expression as the condition for oscillation:

$$Z_0 = \frac{z_1 z_2 z_3 z_4}{(q_1 - z_1)q_2(q_3 - z_4) + z_1 z_2 (q_3 - z_4) + z_3 z_4 (q_1 - z_1) - z_2 z_3 (z_1 + z_4)} \quad (12)$$

In this equation the q's are used to save space and the gen-

In (12) all the a's are positive and then the apparent

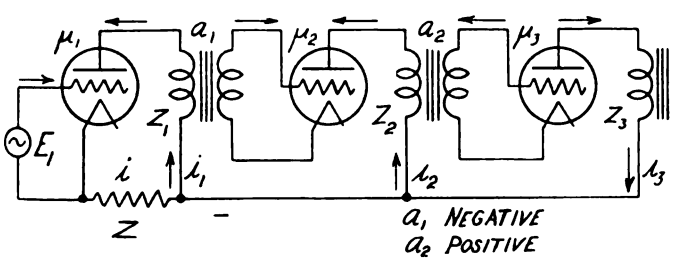

Figure 4d

sign of Z_0 is positive. It is quite possible, therefore, that this represents an unstable connection. If the sign of any one of the a's is changed the apparent sign of Z_0 becomes negative, because the terms containing the a's are dominant.

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But changing the sign of one ratio does not necessarily put Z_0 in the second or the third quadrant for all frequencies. It does, however, shift the frequency band in which oscillation might, and in which regeneration will, occur. It will be noted that the middle transformer occupies a somewhat symmetrical position in the equation as well as in the circuit. It is not affected by end conditions like the other two. The other two transformers produce similar effects on the amplification, and differences are entirely due to differences in tube and load characteristics. Therefore there is very little difference produced by changing the signs of the ratios

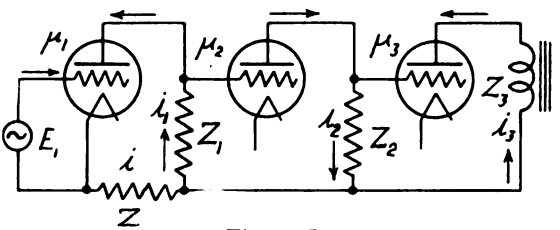

Figure 5

of either. It will also be observed that when all the *a's* are negative, all the terms in the denominator of (12) are negative, which also makes the apparent sign of Z_0 negative. Therefore, one might surmise that this represents the most stable connection. However, when all the signs are the same the absolute value of Z_0 is least, and this would threaten the stability of the circuit in case the phase relations of the impedances should project Z_0 into the first or the fourth quadrant.

Three-Tube, Direct-Coupled Amplifier

The determinant for a three-tube direct-coupled amplifier such as that shown in Fig. 5 takes the form

$$\begin{vmatrix} z_1 & 0 & 0 & Z \\ \mu_2 Z_1 & z_2 & 0 & Y_2 Z \\ 0 & \mu_3 Z_2 & z_3 & Y_3 Z \\ -1 & -1 & -1 & 1 \end{vmatrix} \qquad (13)$$

in which $Y_2=(\mu_2+1)$ and $Y_3=(\mu_3+1)$. When written out this becomes

$$D=z_1z_2z_3-Z[(R_1\mu_2+z_1)\ (\mu_3Z_2-z_3)-z_2(z_3+Y_3z_1)] \qquad (13b)$$

The condition for oscillation in this circuit is, therefore,

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$$Z_o = \frac{z_1 z_2 z_3}{(R_1\mu_2 + z_1)(\mu_3 Z_2 - z_3) - z_2(z_3 + Y_3 Z_1)} \qquad (14)$$

It appears from this equation that a three-tube, direct-coupled amplifier is likely to oscillate. The apparent sign of Z_o is always positive because the negative terms in the denominator are small compared with the positive terms. It is only in case the angles of the various impedances should be such as to project Z_o into the second or the third quadrant that the amplification would be decreased by the common impedance.

Since resistance-coupled amplifiers of odd stages often oscillate, it will be interesting to substitute typical values in (14).

Let $z_1 = z_2 = 1.4 \times 10^5$ ohms, $z_3 = 8{,}000 + j5\omega$, $R_1 = 40{,}000$ ohms, $\mu_2 = 20$, $\mu_3 = 6$, and $Z_2 = 10^5$ ohms. With these values Z_o becomes

$$Z_o = \frac{8{,}000 + j5\omega}{21.36 - j2.755 \times 10^{-4}\omega}.$$

As long as the reactance of z_3 is positive Z_o can never lie in the fourth quadrant. At zero frequency it lies along the axis of reals, and as the frequency increases it advances into the first quadrant. At about 1800 cycles it passes into the second quadrant. Hence the region in which there will be augmented amplification is from zero to 1800 cycles, and for this the reactance of the common impedance must be positive. Inductance is required in series with the common resistance. The amount of inductance needed for oscillation at any frequency may be calculated. For instance, at 1000 radians the critical value of Z is $370 + j238$, and the required inductance is 238 millihenrys. The resistance must also be 370 ohms, or there will be no oscillation. At 10,000 radians the impedance required is $71.6 + j2350$. Thus the required inductance has decreased slightly and the required resistance has been reduced to about a fifth of its former value. The required conditions at both of these frequencies are likely to be met in a battery eliminator that has been inadequately by-passed. And the required resistance may be encountered in an ordinary "B" battery which is still in a comparatively good condition.

If there is an appreciable capacity across the load impedance, Z_o will turn around and pass into the fourth quadrant instead of passing into the second for the higher frequencies. The circuit might then oscillate when there is a condenser across the resistance of the "B" battery. The

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frequency at which the load impedance is in parallel resonance is not likely to be a critical one as far as oscillation is concerned, because the required resistance is very high at that point, although the required reactance is negative and very small.

Four-Tube, Direct-Coupled Amplifier

The equation for Z_0 in a four-tube, direct-coupled amplifier contains many terms of different signs and of uncertain magnitude. It is difficult to interpret without the substitution of actual values. It may easily be obtained by adding another row and another column to the determinant (13), and it is not necessary to give it here. It indicates that the four-tube circuit is more stable than the three-tube circuit. If the added stage is resistance-coupled, and if the constants in this stage are the same as those used in equation (14), the complete expression for the four-tube circuit reduces to

$$Z_0 = \frac{-(8{,}000 + j5\omega)}{328 - j2.185 \times 10^{-3}}.$$

From this equation it is obvious that the circuit represented cannot oscillate until the frequency is high enough to advance Z_0 into the fourth quadrant. This occurs at 15,500 radians, or at 2,470 cycles. For higher frequencies the amplification will be increased by the common impedance provided its reactance is negative. At 10,000 cycles the angle of Z_0 is — 68 deg. 40 min. and the impedance for oscillation is $330 - j830$. At that frequency this impedance is given by a resistance of 2,420 ohms shunted by a condenser of 0.0165 microfarad. This combination is likely to occur in practice.

By tracing the currents in the plate circuits of the amplifier it is clear why the circuit cannot oscillate at low frequencies. Currents i_1 and i_3 are in phase with each other and currents i_2 and i_4 are opposed to them. The sum of the even currents is always greater than the sum of the odd. Consequently, there is more current trying to reduce the amplification than to increase it. But as the frequency increases, less and less of i_4 will be in phase with i_2. There will be one frequency above which the sum of i_2 and the in-phase component of i_4 will be less than i_3. Then the circuit will regenerate. Capacity across the load impedance will prevent i_4 from getting far out of phase with i_2, and consequently this would act as a stabilizer of the amplifier.

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Experimental Evidence in Support of Theory

There is some evidence in support of the theory advanced in this paper. In the first place it is a well known fact that broadcast receivers become noisy during the last stages of life of the "B" battery when its resistance is high, and that the amplification becomes noticeably uneven. Many have also experienced oscillation in sets served by "B" battery eliminators, particularly in three-tube, resistance-coupled circuits. In one case such a set oscillated at a very low frequency when served by a certain eliminator, and it would not stop oscillating until a condenser of thirteen microfarads was put across the output of the filter. Another receiver which had two stages of resistance coupling and one stage of transformer proved to be a terrific oscillator on an ordinary "B" battery. The amplification constants of the tubes in this circuit were in order of occurrence 20, 8 and 6, and the transformer ratio was also six. The equation applicable to the case showed that reversing the ratio of the transformer should have little effect on the behavior of the circuit. Actually it caused only change in pitch. Change of tubes had similar effect.

In an effort to salvage the preceding amplifier it was changed to a transformer, resistance, transformer-coupled circuit. The tubes now used had amplification constants of eight, eight and six, while the transformer ratios were three and two. But this set oscillated, too, and it had three modes of oscillation, one of which was at a super-audible frequency. There was one combination of transformer ratios which gave a stable connection, and this was later used for testing the theory.

When a fresh "B" battery was used, the circuit worked satisfactorily on all combinations of ratios, but it began to oscillate when a resistance of six ohms was inserted in series with the battery. With ten ohms in series the circuit was in an uproar. When a semi-exhausted "B" batteries was used the circuit oscillated violently, and no shunt condenser up to four microfarads had any perceptible effect on the intensity.

With this circuit the following experiment was tried. An extra "B" battery was provided, and the circuit was so arranged that the detector tube could be put on either the common battery or on the separate. When the detector was on the separate battery the circuit did not oscillate even

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when there was a fairly high resistance in series with either of the batteries. But when the detector was put on the common battery the circuit oscillated when there was only a small amount of resistance in series with the battery. The resistance in the common battery was then adjusted until the circuit just stopped oscillating, and the detector was alternately put on the common and on the separate batteries. When on the common, the volume was fully 100 per cent. greater than it was when on the separate. The stable connection was then arranged and the same comparison made. This time the volume was down about 50 per cent. when the detector was on the common battery. Although the voltage on the amplifier was 135 volts, the common portion was only 45 volts, that is, common to all the tubes. When more cells were included in the common battery the tendency to oscillate increased, which was partly due to increased amplification and partly to increased common impedance.

The equation applicable to the above circuit is

$$Z_0 = \frac{z_1 z_2 z_3 z_4}{\mu_3 Z_2 (q_1 - z_1)(q_4 - z_4) + z_3 z_4 (q_1 - z_1) - z_1 z_2 (\mu_3 + 1)(q_4 - z_4) - z_2 z_3 (z_1 + z_4)} \quad (15)$$

in which $q_1 = a_1 Z_1 \mu_2$ and $q_4 = a_2 Z_3 \mu_4$.

From equation (15) one would expect the circuit to be oscillatory when both transformer ratios have the same sign, either positive or negative, and also when a_2 is negative and a_1 is positive. The circuit appears to be most stable when a_1 is negative and a_2 is positive, because then most of the terms in the denominator of (15) are negative.

Methods of Avoiding Oscillation

Several methods of avoiding oscillations may be suggested. The first step, of course, should be taken when the amplifier is designed. Not all types of circuit are equally subject to oscillation, and only those which are relatively stable should be employed. To avoid oscillations and distortion in a set that is already built, the only thing to do is to reduce the common impedance. This may be done by connecting a very large condenser across the plate voltage supply, or the common impedance; by always employing batteries of low resistance; by using a separate battery for one, or perhaps two, tubes in the circuit, one of which

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should be the first tube entering into the amplifier. Another way of reducing the trouble is to employ a choke coil and a series condenser so that the a-c. component of the current in the last tube may be shunted around the common impedance, as is done to protect the loud speaker coils from the d-c. component of the plate current.

While the theory of the vanishing denominator explains why a circuit will oscillate at a given frequency when the impedance in common with the plate circuits has the required value, it does not explain why a circuit will oscillate at any frequency with a given value of the common impedance. But this difficulty disappears when it is remembered that nearly all the circuit parameters which enter into the expression for the critical value of the common impedance, as determined by the vanishing denominator, change both with frequency and with current level. The circuit automatically finds the frequency and the current level where the denominator vanishes, and proclaims the discovery.

SUMMARY

It has been pointed out that the common impedance in the plate circuits of two or more tubes in an amplifier will affect the output, and that under certain conditions will cause distortion and oscillation. A theory has been developed to explain this phenomenon. Several typical amplifier circuits have been analyzed with the theory. Some experimental evidence has been adduced in support of the theory. Methods of avoiding oscillation and distortion have been suggested.

SYMBOLS AND NOTATION

a_n, the voltage ratio of the nth transformer in the circuit.

D, the value of the determinant of the system of simultaneous equations pertaining to an amplifier.

E_n, the input voltage to the nth tube in the circuit.

$E_{n+1}=a_n Z_n i_n$, is the input voltage of the $(n+1)st$ tube when preceded by a transformer of ratio a.

$E_{n+1}=Z_n i_n+Zi$, the input voltage to the $(n+1)$st tube when this is coupled to the nth tube by means of an impedance coil or a resistance.

i, the current in the common impedance Z. It is equal to Σi_n.

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i_n, the a-c. component of the current in the plate circuit of the nth tube.

$j=\sqrt{-1}$

M, the complex amplification of the circuit.

μ_n, the amplification constant of the nth tube.

ω, the angular velocity of the current, or $2\pi\times$ frequency.

q_n, abbreviation for $a_n Z_n$.

R_n, the internal plate resistance of the nth tube (a-c.)

$Y_n=(\mu_{n+1})$.

Z, the common impedance in the plate circuits of all the tubes.

Z_o, the value of Z which makes $D=O$, or which makes the circuit oscillate.

Z_n, the load impedance in the nth tube, exclusive of Z.

z_n, the total impedance in the plate circuit of the nth tube, also exclusive of Z, or $z_n=R_n+Z_n$.

Ψ_o, angle of Z_o $\quad$ θ_n, angle of z_n.

Ψ, angle of Z $\quad$ Φ_n, angle of Z_n.

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SOME POSSIBILITIES AND LIMITATIONS IN COMMON FREQUENCY BROADCASTING

By

De Loss K. Martin,[1] Glenn D. Gillett,[1]
Isabel S. Bemis[1]

The congestion in the broadcasting frequency range is so great that consideration has naturally been given by various engineers to the possibilities of different stations using the same frequency for broadcasting. It is hoped that the following statements may be of some interest in the consideration of this question.

In common frequency broadcasting there are two cases which have to be considered as they are somewhat different in their requirements. The first is the case where two or more stations attempt to use the same frequency with their own separate programs. The second is the case where two or more stations attempt to use the same frequency for sending out a common program which is transmitted to them from a single source.

At the present time, in both cases a limitation on this type of broadcasting is the audible beat notes which are set up between the carriers of the different stations. In the first case if means are taken to keep the carrier frequencies of the different stations closely alike (say within a relatively few cycles), then the limitation becomes that of crosstalk between the separate programs. From the data given in the paper it would appear that stations separated by distances of no more than three or four hundred miles from each other could send out their separate programs in this way provided each station confined the field to which it catered to an area immediately surrounding the stations within a radius of perhaps twenty-five miles or less.

In the second case, if the carrier frequencies are held

[1]Department of Development and Research American Telephone and Telegraph Company.

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closely alike, the nature of the interference becomes somewhat different since each station is transmitting the same program. As in case one, each station, of course, could be satisfactorily heard in a small area immediately surrounding it. This area might be somewhat larger than that of case one since the interference limitations are probably not as exacting. However, in the broader areas between the stations, the wave interferences between the various stations become important and the paper gives some discussion of this fact. The work on which the paper is based was not carried, however, to the point where the limitations set up by such interferences can be predicted.

It is entirely outside the scope of this paper to attempt to draw any conclusions as to the practical force of the limitations imposed by the use of common frequencies in broadcasting or as to what use, if any, should be made of such common frequency broadcasting. The subject merits careful consideration in any case in which stations are sufficiently close in frequency and in distance so that beat notes between their carriers become objectionable.

Interference Between Stations

To consider the matter somewhat more in detail, in both cases above, it is evident that if the carrier frequencies of the various stations are held to a common value and not allowed to depart from it by an amount which will produce audible frequency beat notes between the various carriers, such beat notes cannot exist. In other words, it would seem possible by more accurate frequency adjustment to avoid the principal difficulty that has been met with in the past in endeavoring to operate two stations on the same nominal frequency with any other than a very wide geographical separation. If in this way interfering noises due to the interaction of two carriers are avoided, the next difficulty would appear to be that of interference of one program to another.

Since with two stations with different programs on the same frequency it is impossible to separate the programs by any selective tuning arrangements, the relative signal strengths of different stations at a given point would determine whether the program from any one of them could be satisfactorily received. In order to obtain satisfactory re-

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ception, the signals from the desired station must necessarily be of much greater amplitude than those from the interfering stations.

The range of a station would, for satisfactory results, be limited to that area within which it was capable of producing signals which were stronger than the signals of any other station on the same frequency by whatever amount might be felt to be necessary for satisfactory freedom from crosstalk interference.

If carriers of the various stations differed by, perhaps, a few cycles per second, it is obvious that when the points of the carrier field strengths are not widely different in amplitude, the net strength of the received carrier signal would pulsate or flutter. Within the range of a station as determined by crosstalk considerations, it is not to be expected that this flutter would be of any material importance, since the amount of flutter or variation is determined by the adding and subtracting of the two carrier signals and if one of these is a great deal less than the other, the fluctuation would probably be imperceptible. It can be shown by mathematical analysis that even where the local signal amplitude flutters by only an extremely small amount the amplitude of the distant signal will flutter between zero and its maximum amplitude synchronously with the local signal. The complete flutter of the distant signal would probably so entirely destroy its identity that the crosstalk from it would be a substantially unintelligible noise. As to what the condition would be outside of the limited crosstalk free service areas of the stations, it is impossible to predict with assurance, since almost no experimental data are available. It seems likely, however, that the program crosstalk between stations and the fluttering of signal amplitude, combined with ordinary fading, would cause a very unsatisfactory situation.

Now consider the case where a number of stations on the same carrier frequency transmit a *common program.* The principal difference between this and the case just discussed is that the same program is transmitted from all stations in unison and it seems possible that more interference between programs could be tolerated due to the fact that no audible signal would be present during silent intervals of the program. If this is true, satisfactory reception might be obtained in areas where one signal dominated the

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others by a lesser amount than is required for the case where different programs are transmitted. If the program was distributed to the various stations over long wire telephone circuits, or in some other way which might, due to transmission delay, cause an appreciable lack of simultaneity in the transmission of the program from different stations, it might be that various signals would arrive at a given receiving station so much out of step with each other as to be nearly as objectionable as would be the case with entirely different programs.

It is of interest to consider what might occur if the various carrier frequencies of stations transmitting the same program were held to exactly the same value with a high degree of precision. This would mean that instead of having a flutter, the frequency differences would be so small that the flutter would be slowed down to something which would be more like fading. The interfering signals from the distant station would then become more intelligible and less objectionable. The extent to which this would increase the satisfactory range of the station before other limiting factors became of importance, can be determined only by experiment. It seems certain, however, that at points which are well away from any one of the stations, that is, where the signals from two or more of them are of approximately the same amplitude, continual fading variation of the signals, due to these slight frequency changes, would exist. Superimposed upon this there would no doubt be variations due to ordinary fading of the individual signals themselves. In addition not only would carrier frequency signals of the various stations interact in this way, but each individual side-band frequency would be similarly affected and no two frequencies would be necessarily affected in the same way at the same instant. In other words, the phenomena would be comparable with those which result from selective fading, and selective fading is known to cause damage to quality. It is a question which can be determined only by experiment whether this combination of variations would be so annoying as to completely prohibit satisfactory reception, or whether, on the other hand, the averaging effect of all the variations might produce a resulting signal, which would have no more annoying net variation than would the signals received from any one of the stations alone, presuming the others to be silent.

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Frequency Control

In the practical case of two or more common frequency broadcast stations the carrier frequency could be determined by a single centrally located control-oscillator or some or all of the stations could have independent oscillators carefully standardized against each other.

In the case where a central control oscillator is to be used, it is desirable that the frequency of the oscillator be relatively low. The carrier frequency of the broadcast stations could be derived as a harmonic of this signal after it had been sent to the stations over wire circuits.

As the transmission characteristics of wire circuits change with the temperature and weather conditions the phase relation of the control signal at the sending and receiving terminals of the wire circuit would vary. The changes in phase of the control signal would be relatively small, but they would be multiplied in magnitude in the radio signal in proportion to the order of the harmonic used to obtain the radio carrier signal. For example, suppose a 1000 kc. carrier signal is to be obtained from a 1 kc. control signal sent over a wire line, then the phase changes in the 1 kc. control signal would be magnified a thousandfold in the carrier signal.

The significance of such phase changes is that while they are going on they are in effect frequency differences. Thus if the phase of one station changes with reference to that of another station at a rate of 360° per second the two station frequencies actually differ at that time by one cycle per second.

It seems likely that a central control-oscillator system would be necessary if it were desired to attain the greatest possible precision in synchronizing the various stations. On the other hand it is probable that independent oscillators, for instance those of the piezo-electric crystal type, could be standardized with sufficient exactness to avoid audible beat notes between stations.

Quite aside from other limitations there must be practical limits to the number of stations transmitting on one common frequency, due to the fact that cities in which stations might be located are not spaced at uniform intervals over the country and also to the fact that the power radiated from the various stations might not be equal.

It is evident that the existence of one common frequency

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broadcasting channel used simultaneously by a number of stations does not exclude the possibility of there being other common frequency channels as well, nor does it in any way alter the necessity for adequate frequency spacing between stations serving the same area on different channels.

Experimental Data

To illustrate a little more concretely the nature of the problems involved in common frequency broadcasting, use is made of the data obtained from a preliminary survey which was made during the spring of 1926, of the signal

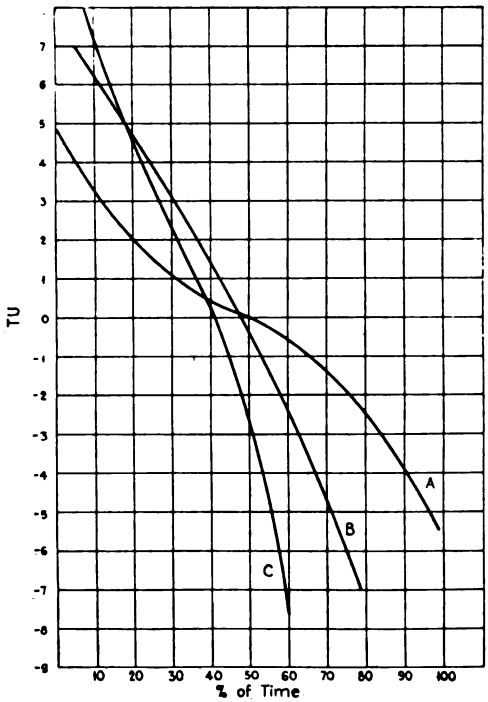

Figure 1. Variation in amplitude of audio-signal from the mean value caused by fading, in terms of percentage of the total observing time the signal exceeded these levels. Station 2XB.

A. Observed at Philadelphia, Pa.
B. Observed at Baltimore, Md.
C Observed at Washington, D. C.

distribution from stations 2XB New York City and WCAP Washington, D. C. while these stations were transmitting on their normal frequencies. This permitted estimating the

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effective ranges that would be obtained from these stations should they operate on a common frequency.

It was found that in the daytime the signal intensity from these stations remained substantially constant. At night graphic records of the variations in the signal intensity due to fading were made for each station at each of the observing points.

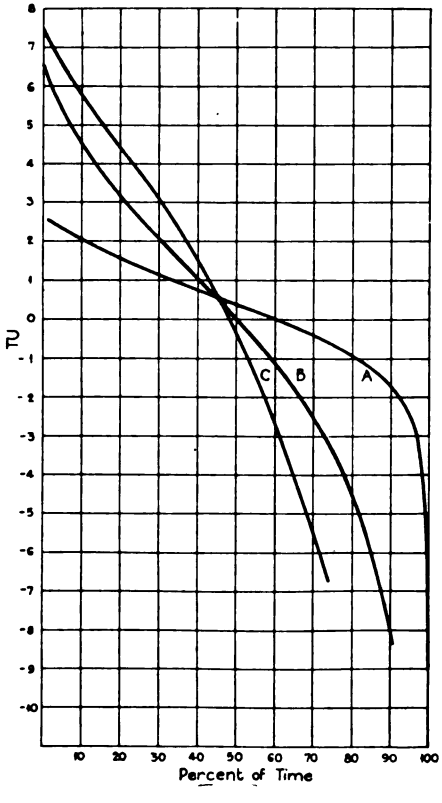

Figure 2. Variation in amplitude of audio-signal from the mean value caused by fading, in terms of the percentage of the total observing time the signal exceeded these levels. Station WCAP.

A. Observed at Baltimore, Md.
B. Observed at Philadelphia, Pa.
C. Observed at New York, N. Y.

From the graphic records it was possible to obtain mean values of the equivalent audio signal intensity and also to obtain an estimate in terms of per cent of the total observing time that the audio signal exceeded values which departed from this mean by any given amount. Figures 1 and

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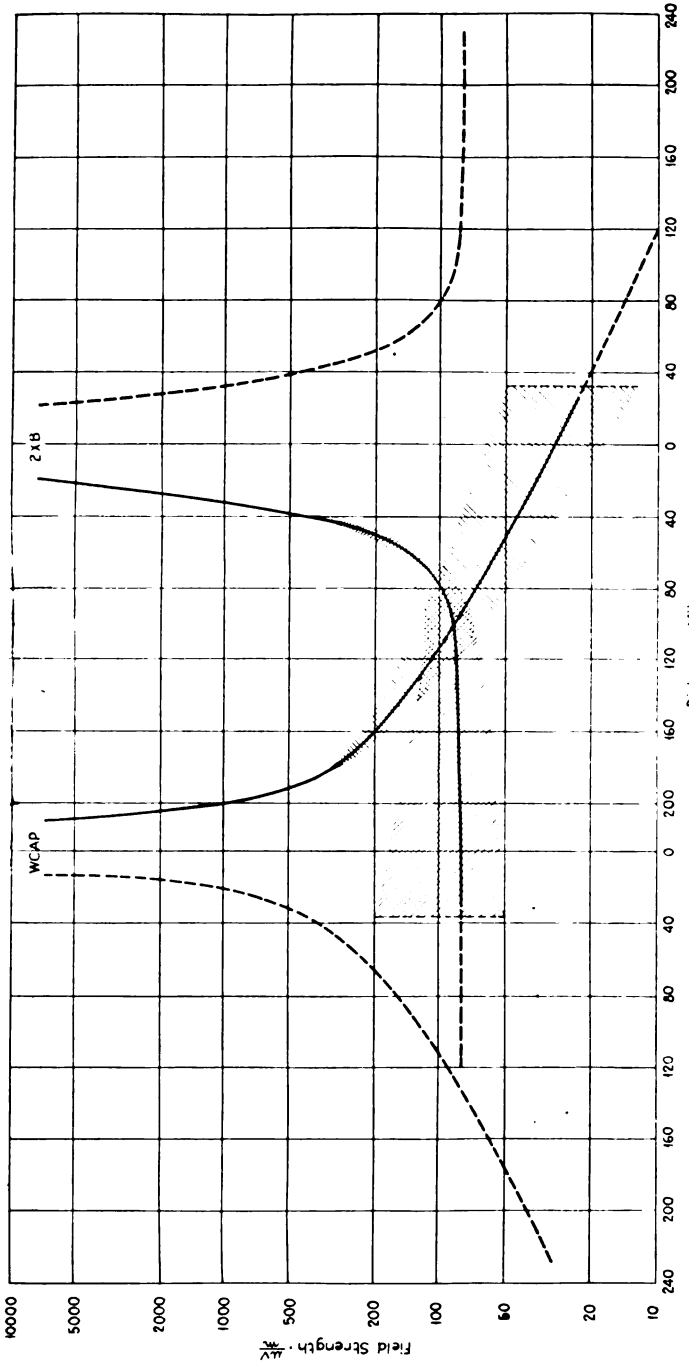

Figure 3. Measured variation of field intensity with distance for 2XB and WCAP. The shaded part indicates roughly the magnitude of the signal variation due to fading. From these curves an estimate of program interference may be obtained.

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2 illustrate the characteristics of the variation in the signals recorded on these nights. The curves are representative of the variations taken from several graphic records, each extending over a period of from 10 to 20 minutes and taken at intervals of an hour or so. A very interesting fact brought out by the data on field intensity measurements is that the values of the steady daytime signal intensity and the mean of the signal intensity measured at night are substantially equal. This is true except for one set of observations on the 2XB signal made at Washington.

In Figure 3 the daytime signal field intensities are plotted for both stations against distance, and the observed

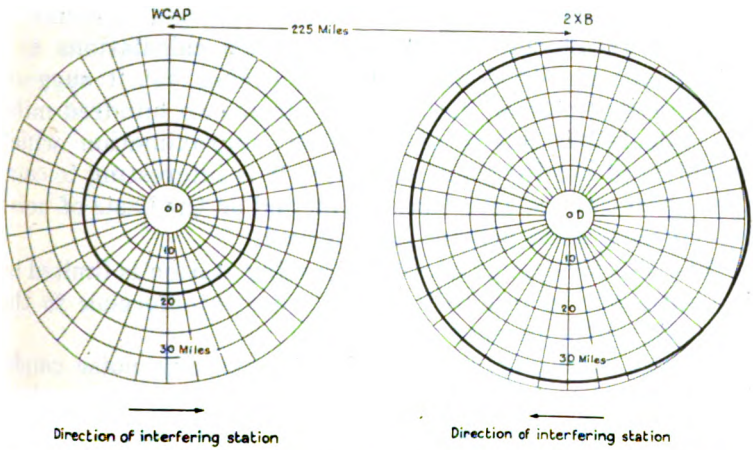

Figure 4. Effective ranges of 2XB and WCAP as they would be limited by program interference if both stations transmitted on the same frequency. Based on measurements made at night during May, 1926.

ranges of variation in field intensity are shown approximately by the limits of the cross-hatched areas above and below the curves of the daytime signal.

For the purposes of making an estimate of the effective range areas of these two stations if they were to be operated on a common frequency, it was assumed that the signal distribution was symmetrical about the stations as shown by the curves. This assumption is only partly true but is necessary to simplify consideration of the analysis.

From the data given in Figure 3 and on the basis of a ratio of about 30 to 1 between the desired carrier signal and

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the interfering carrier signal (a ratio which is perhaps too lenient) the range areas shown in Figure 4 were obtained. These range limits correspond to those that would be obtained during the day and under average night conditions.

However, in this experimental case the upper limit of the fading night-time signals differs so little from their mean value that the range is not materially reduced by the fading.

The range shown in Figure 4 for 2XB was compared with theoretical range calculated on the simple assumption that the field strength from both stations varied inversely with the distance. It was found that this calculated range had a radius about two-thirds as large as that derived from the experimental data. This is not sufficient evidence to warrant recommending the use of such calculations as a conservative method of estimating ranges but it suggests that for interstation distances of about two hundred miles the calculation might give a first approximation which would be of use for some purposes. It is interesting to compare these ranges with those estimated on the basis of noise interference as at present experienced.

The effective range of a broadcasting station is limited by the ratio of the intensity of the noise interference to the intensity of the desired signal.

It has been suggested[1] that with existing noise conditions, the outer boundary of the service range of a station lies between one and ten millivolts per meter and in general is nearer the latter value. From the field intensity contour maps of 2XB and WCAP which have been published,[2] the ranges for different field strengths can be determined. It is found that even for one millivolt per meter the ranges are of the same order of magnitudes as those shown on Figure 4.

We do not wish to infer from this comparison that the ranges of 2XB and WCAP would not have been materially affected if they had been placed on a common frequency to broadcast a common program. The assumptions underlying the above comparison are not sufficiently well supported to justify such a conclusion. On the other hand the fact that it comes out as a comparison rather than a contrast indicates that the subject is one which merits further and more exhaustive investigation. It should also be borne in mind that outside the ranges as determined for the common frequency case, reception of either station might be hopeless at all

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times, whereas with the stations on different frequencies it is known that they are frequently received with material satisfaction at distant points.

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THE INSULATION OF A GUYED MAST*

BY
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(Federal Telegraph Company, Palo Alto, California.)

In planning to use guyed masts for supporting the antennas of high power transmitting stations, the question arises as to the advisability of insulating the masts and guys. Under certain conditions this will increase the effective height and efficiency of the antenna giving a power saving that is worth more than the cost of the insulation and the increase in mechanical hazard. Under other conditions the increase in antenna efficiency may be insufficient to make the insulation worth while.

Among the factors affecting the antenna efficiency some, such as the resistance of the antenna, can be improved after the antenna has been built. The mast and guy insulation, however, must be decided before erecting the masts and it is important to determine beforehand, (a) the most efficient arrangement of insulation, (b) the maximum voltage duties to be expected on the insulators, and (c) the economic value of the insulation. No fixed rules can be given in this regard since each design of antenna and mast system offers a different problem. An attempt will be made, however, to point out in this paper, with the aid of a characteristic example, certain fundamental considerations and experimental methods that will assist in the economical insulation of guyed masts.

THEORETICAL CONSIDERATIONS

A general idea of the effect of mast and guy insulation on antenna effective height can be obtained by considering the properties of an ideal antenna. If a large flat antenna plate of fixed area float in air above a flat ground plate of infinite extent, the capacitance of the antenna will depend

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upon the height of the plate above ground, or its effective height. The greater the effective height, the less the capacitance and, contrarywise, the greater the capacitance, the less the effective height.

The characteristics of the electrostatic field around such an antenna are shown in Fig. 1. This is a vertical section

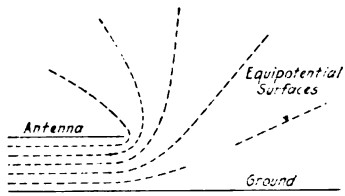

Figure 1—Ideal Antenna

through one side of the antenna plate and gives the approximate traces of the equipotential surfaces. It is seen that these surfaces are close together between the antenna plate and ground, but spread out beyond the edge of the plate.

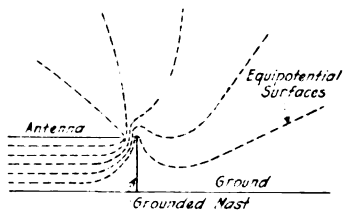

Figure 2—Antenna with Grounded Self-supporting Mast

If a grounded self-supporting mast be moved up close to the edge of the antenna, the equipotential surfaces will be distorted in the manner shown in Fig. 2. This distortion occurs only at the mast, the equipotential surfaces at some distance in front and back of the mast remaining as in Fig.

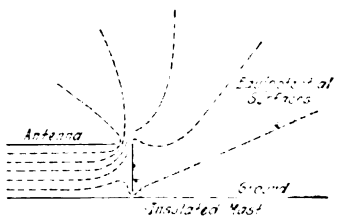

Figure 3—Antenna with Insulated Self-supporting Mast

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1. The mast, therefore, has the effect of crowding the equipotential surfaces between its top and the edge of the antenna plate.

With the mast insulated at the ground, the distortion is not so great (see Fig. 3), the insulation at the base of the mast allowing one of the equipotential surfaces to pass underneath the mast.

The effect of crowding the equipotential surfaces is to increase the electrostatic flux and hence the capacitance

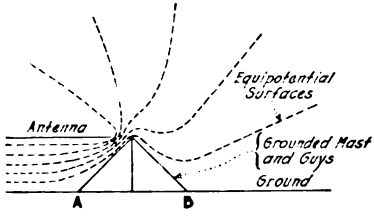

Figure 4—Antenna with Grounded Guyed Mast.

between the antenna and ground. The grounded mast of Fig. 2, therefore, increases the total capacitance of the antenna and decreases its effective height. With the insulated mast of Fig. 3, there is less crowding of the equipotential surfaces, less increase in capacitance, and less decrease in effective height. If the mast were also insulated at its middle point, an approach would be made to the ideal condition shown in Fig. 1.

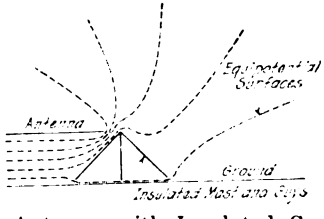

Figure 5—Antenna with Insulated Guyed Mast

The use of guys on a mast decreases the effective height still further. Fig. 4 shows the effect of uninsulated guys on a grounded mast. It is seen that the equipotential surfaces are crowded more than in Fig. 2.

If, in the case of Fig. 4, Guy A were insulated at a point near the ground, none of the equipotential surfaces could pass in through the insulator unless they also came out

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through it. The voltage duty on the insulator would, therefore, be zero. Insulating all the rest of the guys and the mast near the ground, would permit some of the surfaces to pass under the entire structure as shown in Fig. 5. The crowding of the surfaces would, however, be greater than in the case of Fig. 3.

With an insulator in each end of Guy A, the equipotential surfaces could pass in through one insulator and out through the other. The more surfaces that would pass through the guy insulators, the less crowded would be the surfaces between the antenna and guy system and the closer would be the approach to the condition of Fig. 2, where there are no guys.

The ideal guy would be one made up entirely of insulation. This would allow the equipotential surfaces to pass through it as though there were no guy and give the maximum value of effective height. Since the greatest possible number of surfaces would pass through the guy insulation the total voltage duty on the guy would be greater than in any case where it is not made up entirely of insulation. For maximum effective height, it is, therefore, necessary to have the total voltage duties on the insulators in each guy a maximum. The voltage duties on the mast insulation should also be kept high for the same reason.

It is not usually practicable to insulate a mast and its guys at very many points, but it is possible to insulate the guys at a few points so as to obtain such a large percentage of the maximum total voltage duty, that the addition of more insulation would cause very little increase in the total voltage duty. This is the economic condition desired in the arrangement of the insulation.

A simple relationship exists that is of great help in studying this problem. It was shown above that with two insulators in Guy A of Fig. 4, the equipotential surfaces can pass in through one insulator and out through the other. Since the same number of surfaces pass through each insulator, the voltage duty on each must be the same. In the same way, if three insulators are placed in the guy a certain number of surfaces pass through one insulator which must pass out through the other two. The voltage duty on the first insulator must then be equal to the sum of the voltage duties on the other two insulators.

Another way of looking at this relationship is to con-

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sider the potential above ground of the insulated section or sections of the guy. With an insulator in each end of a guy on a grounded mast, one side of each insulator will be connected to ground. The other side will be connected to a section of guy cable in which there is practically no potential drop. The sides next to the cable must, therefore, be at the same potential above ground so that the voltage duty on each insulator is the same. When there are three insulators in the guy, one of the insualted guy sections must be at a higher potential above ground than the other. The drop in voltage between the ends of this high-potential section and ground must be the same so that the sum of the insulator voltage duties above it must equal the sum of the insulator voltage duties below it. If then, the voltage duties be given signs according to whether they occur above or below the high-potential section, the algebraic sum of all the voltage duties in the guy must equal zero.

The same relationship occurs when the mast is insulated. In this case, the high-potential section can be the insulated section of the mast or a guy section. In the example shown in Fig. 14, the high-potential sections are in some cases in the guys and in other cases in the mast. In each case, the sum of the insulator voltage duties on one side of the high-potential section are equal and opposite to the sum of the duties on the other side.

The actual voltage duties on the insulators will depend on the position of the guys and masts with reference to the antenna. Insulators in guys close to the antenna will be subjected to high potentials on account of the large number of equipotential surfaces which can pass through them. Those in guys away from the antenna or shielded by other guys, will be subjected to very low potentials. The actual values for a given antenna and mast system can be obtained by measurement on the antenna or on an accurately scaled model. The best arrangement of insulators can best be determined on a model.

Another possible use for mast and guy insulation, is in the control of capacitance currents. It was pointed out above that if Guy A, of Fig. 4, were insulated near the ground, the voltage duty on it would be zero. An insulator at this point, however, would have a high reactance so that any capacitance currents in the guy would flow off to ground through the mast. This arrangement has an advan-

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tage in case a better ground connection is available at the mast than at the guy. Its value would depend upon the type of ground system used and can be estimated sufficiently accurately before the masts are erected.

Experimental Determinations

In order to show how these theoretical considerations work out in a practical case, the results of certain voltage duty measurements made on the model of a typical high-power transmitting antenna will be given. These measurements were mainly to determine the best location for the insulators and the voltage duties to be expected. They did not include any effective height measurements which would permit a determination of the economic value of the insula-

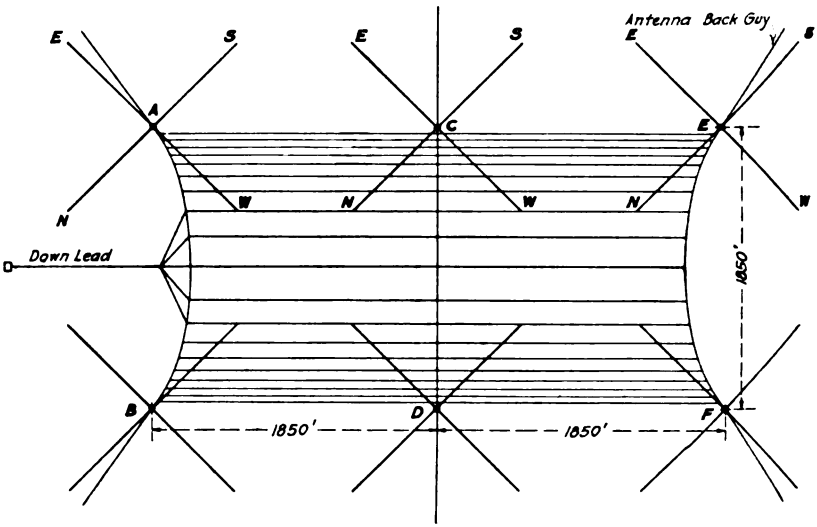

Figure 6—Plan of Antenna

tion nor measurements which would indicate its value in the control of capacitance currents.

The voltage duty measurements were made on a 1-50th-scale model of the antenna shown in Fig. 6. This antenna was designed for use with six steel masts each 1000 feet high and guyed in four directions in the manner shown in Fig. 7.

The antenna model was built in an open field as accurately to scale as possible. The masts were 20-ft. lengths of

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1¼-inch galvanized iron pipe held in place by guys of No. 12 iron wire, stretched tight with turnbuckles. They were also insulated from ground by means of small porcelain tubes about 1½ inches long, jumpers being provided for shorting this insulation when desired. Each guy anchor was well grounded by soldered connections to a metallic stake and a common ground strip.

The guy insulators consisted of short lengths of ½ by ¼-inch bakelite, with hooks inserted in each end, the creep-

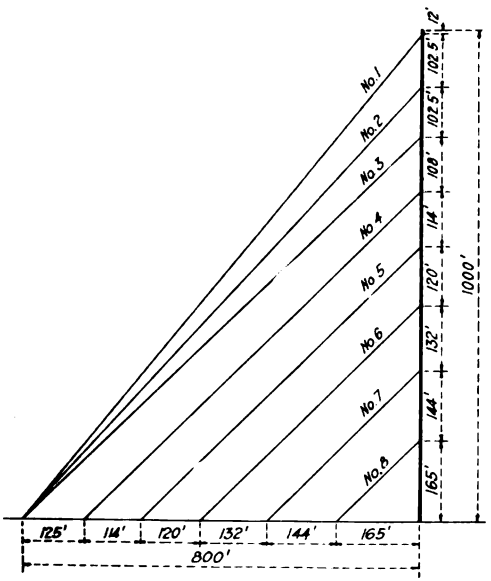

Figure 7—Guy Arrangement for 1,000-Foot Mast

age distance between hooks being one-half inch. They were the same length as the sphere gaps used for measuring the voltage duties, and were placed at every point in the guys where measurements were desired. Any unused insulators were simply short-circuited out with short lengths of spring wire.

The antenna was made of No. 22, B. & S. gage copper wire stretched over catenaries of ⅛-inch diameter flexible iron cable. It was insulated with ⅜-inch diameter pyrex rod insulators having two-inch diameter corona shields spaced to give a flashover distance of three inches.

Voltage duty measurements on the antenna model were made by means of a sphere gap. This was put either across

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or in place of the insulator whose duty was to be measured and adjusted so as to flash over when a high potential was applied to the antenna. Since the flashover voltage of the gap was the insulator duty at the instant of flashover, the ratio between the gap voltage and the antenna voltage required for flashover was the ratio between the duty on the insulator and the antenna voltage. This ratio would be the same for both the model and full sized antennas. By assuming a given antenna potential (150 kilovolts was used in these tests) it was possible to reduce all readings to a common basis.

In the preliminary tests, radio frequency potentials were applied to the antenna, flashover being detected visually through a surveyor's transit. Voltage control at the frequency used (35,000 cycles) was not simple so that the readings were erratic and not reliable. Extensive tests then showed that 60-cycle potentials could be used with greater constancy in results and in agreement within 0.6 per cent., with observations made over a wide range of radio frequencies. It was not possible to see the flashovers, however, so that it was necessary to listen for them. When the gap was used close to the ground, the observer could get within easy hearing distance but for high positions a long rubber listening hose had to be attached to the gap. Some measurements have since been made on larger models where the gaps were so high off the ground as to make a hose impracticable. In this case a short wave loop receiver was used which tuned quite sharply to the flashover at about 40 meters.

The desired 60-cycle potentials were obtained with a 110/33,000-volt potential transformer using a voltage control rheostat in the 110-volt supply circuit. Meters for measuring the antenna voltage were also placed in this circuit since tests showed that the voltage ratio on the transformer remained constant under the conditions of these tests. Antenna potentials above 30,000 volts could not be used on account of the antenna model going into corona.

The sphere gap adopted after considerable development is shown in Fig. 8. It consisted of a piece of natural bakelite tubing with brass inserts in each end through which the spheres (turned from ¼-inch brass screws) were inserted. Lock nuts were used to fix the length of the gap. On one side of the bakelite tube was inserted a smaller bake-

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lite tube for attaching the rubber listening hose. On the other side a hole was left so as to expose the gap to the atmosphere. The distance between hooks on this unit was the same as on the guy insulators.

Three of these sphere gaps were used, each being set with a different gap length. Wherever possible the largest gap was used since this was the easiest to hear and gave the most consistent results. The smallest gap was used where the insulator duty was low. In some cases the duties were so very low that it was impossible to get a reading even with this gap.

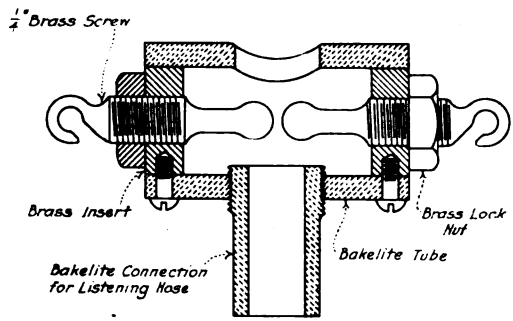

Figure 8—Sphere Gap for Voltage Duty Measurements

After a set of observations had been made on the model guys, the gaps were calibrated directly across the terminals of the transformer. The best results were obtained when the gaps were polished and calibrated frequently.

Arranging the Insulators

To get a preliminary idea as to the best locations for the insulators in the mast guys, measurements were made on a single guy attached to the top of Mast C and extending under the antenna at an angle of 90 deg. to its edge. The the guy, moving the second one up the guy increased the results of the tests with two insulators in this guy are shown in Fig. 9. With one insulator fixed at the bottom of duty on each of them, reaching a maximum relative value of 6.4 near the top. Leaving the second insulator close to the mast, the first one was moved up the guy. This increased the duty on each one to 9 at a point about half way up the guy. Above this point the duties dropped off and be-

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came zero when the two insulators were again merged into one.

This test showed that the best electrical arrangement for two insulators in a single guy is to have one near the mast and the other about the middle of the guy. From a practical standpoint, however, it is preferable not to have an insulator where it will increase the tension on the guy due to its weight or to wind pressure on its surface. It

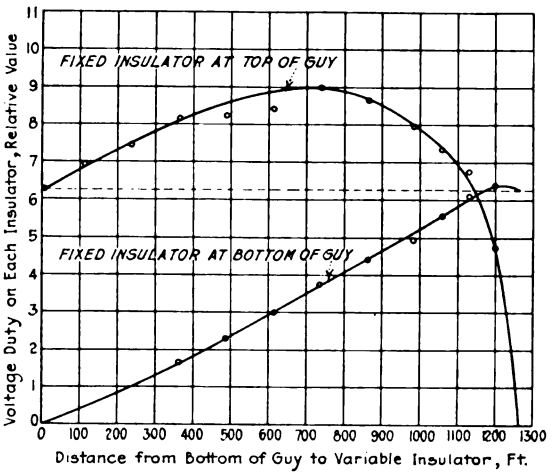

Figure 9—Curves Determining the Best Positions for Two Insulators in Single No. 1 Guy Placed at 90 degrees with Edge of Antenna on Mast C.

was, therefore, considered advisable to keep the insulators close to the ends of the guys where maintenance would be easier.

Further tests showed that the addition of a third insulator in the guy increased the total voltage duty enough to make its use worth considering. Fig. 10 shows the results of tests made to determine the best position for a third insulator. One insulator was fixed at the bottom of the guy, another one was fixed at the top, while the third one was moved between the two. The curves show the variation in voltage duty on the three insulators. They also illustrate the summation principle outlined above. Up to the point where the middle insulator has zero duty, the duty on the top insulator is numerically equal to the sum of the duties on the other two insulators. Above this point, the

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duty on the bottom insulator is equal to the sum of the other two. Although the curve for the middle insulator was obtained from the other curves by difference, enough check measurements were made to make sure that this procedure is correct.

The fact that the duty on the top insulator for most

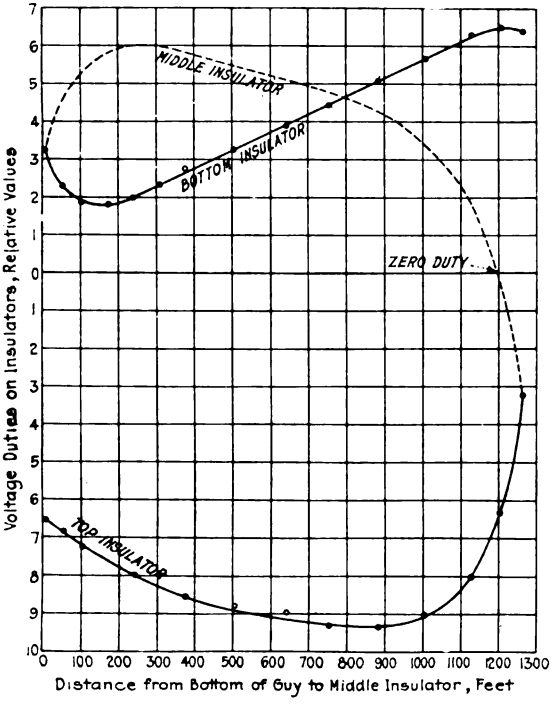
MIDDLE INSULATOR
BOTTOM INSULATOR
ZERO DUTY
TOP INSULATOR
Voltage Duties on Insulators, Relative Values
Distance from Bottom of Guy to Middle Insulator, Feet
0 100 200 300 400 500 600 700 800 900 1000 1100 1200 1300

Figure 10—Curves Determining the Best Position for Third Insulator in Single No. 1 Guy Placed at 90 degrees with Edge of Antenna on Mast C. Insulators Fixed at Top and Bottom. Position of Middle Insulator varied.

cases was equal to the sum of the other two, permits this curve to be used in determining the best location for the middle insulator. The total duty on the guy is twice the duty on the top insulator so that maximum duty on the guy occurs with maximum duty on the top insulator. It is interesting to note that this point of maximum duty occurs when the duty on the bottom and middle insulators is the same.

Results of similar measurements for a more practical case are given in Fig. 11. In this case eight guys were in

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place at a small angle with the edge of the antenna, as position CN of Fig. 6. All the guys were insulated at each end. The position of the middle insulator of the top guy was changed and the voltage duties in that guy measured. The

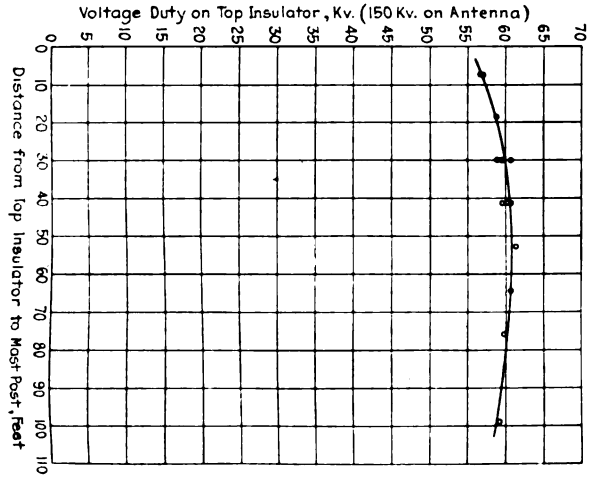

Figure 11—Curve Determining the Best Position for Third Insulator in No. 1 Guy at Position CN with all Guys in Place and Insulated. Top Insulator 42 Feet from Mast. Bottom Insulator 45 Feet from Ground.

curve shows that the maximum guy duty occurred when the middle insulator was about 70 per cent up the guy.

With the same guy arrangement, Fig. 12 shows the results of leaving the bottom and middle insulators fixed and moving the top insulator. It is seen that the position of this insulator is not critical, but that maximum duty on the top insulator occurs when it is about 50 feet away from the mast post. Since the duty on this insulator is half the total guy duty, maximum duty also occurs with the top insulator fifty feet from the mast post. For this test the antenna was the equivalent of about 125 feet from the face of the mast.

Insulator Voltage Duties

As a preliminary study, all guys were insulated at the top and bottom, and voltage duty measurements made at six characteristic guy positions with the masts both grounded and insulated. The results are given in Table 1. As might have been expected, maximum duties occurred on the inside position CN (See Fig. 6.) Minimum duties oc-

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curred in position AE with the masts grounded, and in position EW with the masts insulated.

The duties at positions CN and CE are shown also in Figs. 13 and 14 and have been used for plotting traces of the equipotential surfaces. It will be noted that separate duties are indicated at the bottom of the three upper guys. This is because the guys are insulated separately and not

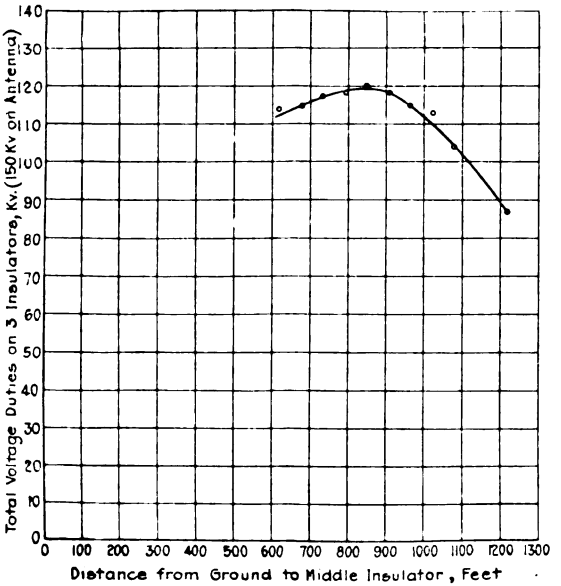

Figure 12—Curve Determining the Best Position for Top Insulator in No. 1 Guy at Position CN with all Guys in Place and Insulated. Middle Insulator 835 Feet from Ground. Bottom Insulator 45 Feet from Ground.

with one insulator as shown. These figures furnish a good check on the theoretical and experimental results previously outlined. In the case of the grounded mast (Fig. 13), four equipotential surfaces enter the guy system at the left through the bottom insulator of the top guy. After spreading out through the guy system, they come together again and pass out through the top insulator in the top guy. On the other side of the mast two of the surfaces enter the guy system at the top and pass out again at the bottom. The fact that the equipotential surfaces on both sides of the guy system bend down in passing through the bottom insulators shows that it would be better to move these in-

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sulators up the guys as was indicated by the curves in Fig. 9.

When the C mast was insulated (see Fig. 14) some of the equipotential surfaces passed underneath the mast as in

TABLE I

Voltage Duties with Insulator at Top and Bottom of Each Guy.

Guy No.	Insulator Position	Kv. Duty on Insulator (150 kv. on Antenna) AN*	AE*	AW*	CN*	CE*	EW*
MASTS GROUNDED							
1	Top or bottom	21.3	12.2	36.4	44.1	23.7	18.3
2	" " "	18.1	10.2	28.9	37.0	20.6	15.5
3	" " "	15.6	8.4	24.4	29.9	17.6	12.4
4	" " "	10.6	5.8	16.7	19.7	13.5	9.5
5	" " "	7.8	4.7	11.6	14.8	10.0	7.3
6	" " "	5.6	3.5(e)	7.6	10.0	7.1	5.1
7	" " "	4.5	2.5(e)	4.5(e)	5.4	4.1	4.0(e)
8	" " "	2.5(e)	1.5(e)	2.5(e)	4.0(e)	2.5(e)	2.5(e)
½	Total Duties	86.0	48.8	132.6	164.9	99.1	74.6
	Total Duties	172.0	97.6	265.2	329.8	198.2	149.2
MASTS INSULATED							
1	Top	3.5	4.5	18.5	18.8	0.5	1.8
	Bottom	23.5	15.5	38.2	46.3	28.0	19.8
2	Top	1.2	5.3	13.1	11.5	2.6	0.2
	Bottom	21.2	14.7	33.1	39.0	24.9	17.8
3	Top	0.8	6.0	8.5	6.5	4.7	2.2
	Bottom	19.2	14.0	28.5	34.0	22.8	15.8
4	Top	5.3	7.7	0.1	3.5	9.5	5.2
	Bottom	14.7	12.3	19.9	24.0	18.0	12.8
5	Top	8.0	8.8	5.5	8.2	11.8	7.6
	Bottom	12.0	11.2	14.5	19.3	15.7	10.4
6	Top	10.2	9.4	8.4	13.0	14.6	9.0
	Bottom	9.8	10.6	11.6	14.5	12.9	9.0
7	Top	12.6	9.5	11.7	16.4	17.5	11.0
	Bottom	7.4	10.5	8.3	11.1	10.0	7.0
8	Top	13.4	9.6	14.0	18.9	19.3	12.6
	Bottom	6.6	10.4	6.0	8.6	8.2	5.4
	Total Duties	169.4	160.0	239.9	293.6	221.0	147.6
Mast Duties: A equals 20.0 kv. C equals 27.5 kv. E equals 18.0 kv.							

(e) equals estimated.

* For notation of guy positions see Fig. 6.

Fig. 5. It would appear, however, that an improvement in the location of the guy insulators could be made in this case. For instance, lowering the top insulators in the five lower guys would cause less distortion of the equipotential surfaces.

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In studying the results given in Table 1, it was seen that the voltage duties on some insulators were so low as to make them uneconomical, and on others were so high as

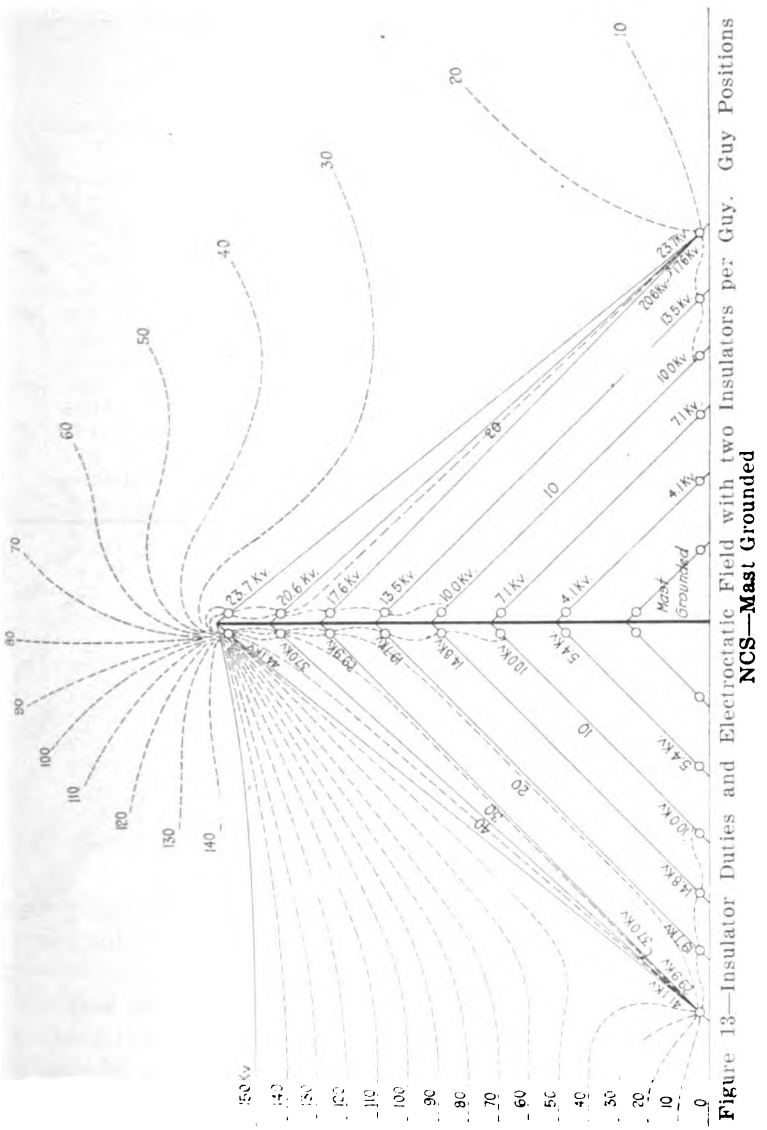

Figure 13—Insulator Duties and Electrostatic Field with two Insulators per Guy. Guy Positions

to make additional insulation advisable. The insulators in the underneath guys of Fig. 13 for example, were practically unused while those in the top guy under the antenna

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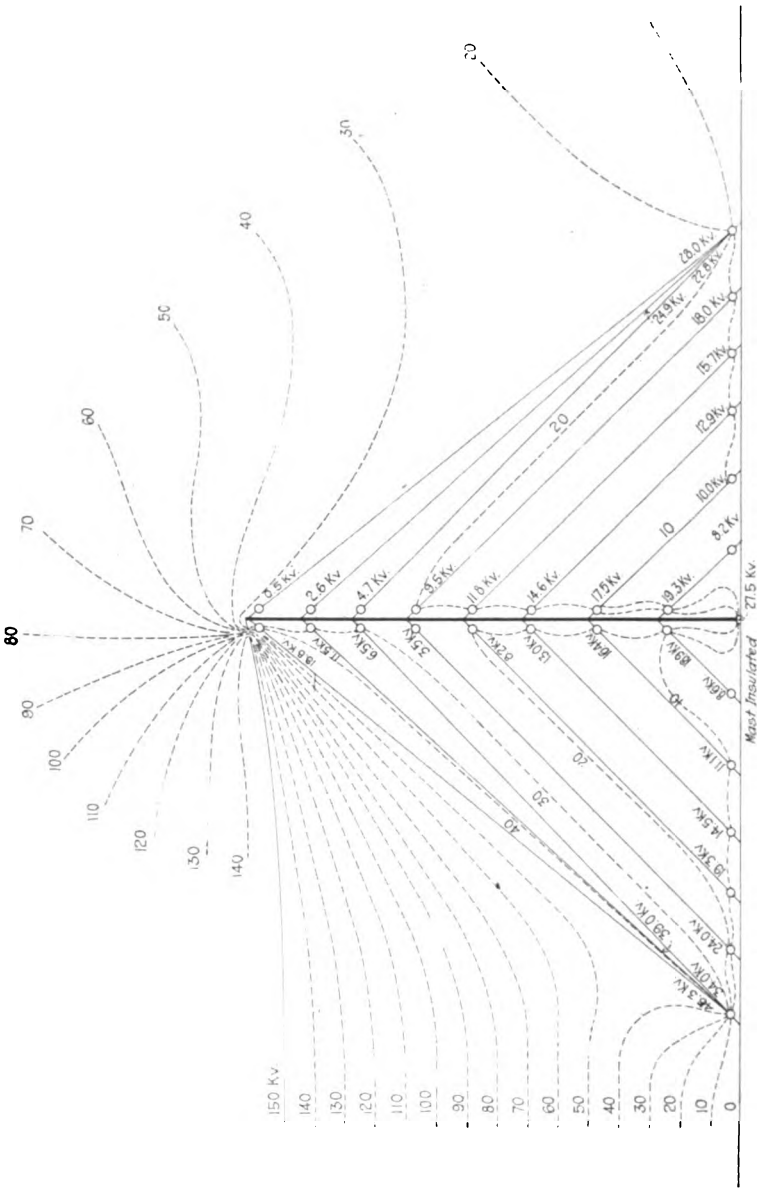

Figure 14—Insulator Duties and Electrostatic Field with 2 Insulators per Guy. Guy Positions NCS—Mast Insulated.

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had very high duties. The question then arose as to where to eliminate insulation and where to increase it. The answer to such a question depends fundamentally on the corresponding change in effective height, but it is also influ-

TABLE II

Voltage Duties with Minimum Total Duty per Guy equals 10 kv.—Masts Grounded.

Guy No.	Insulator Position	Kv. Duty on Insulator (150 kv. on Antenna)					
		AN*	AE*	AW*	CN*	CE*	EW*
1	Top	27.6	12.2	44.7	57.8	31.1	23.3
	Middle	13.8	Out	22.3	28.9	15.5	11.6
	Bottom	13.8	12.3	22.4	28.9	15.6	11.7
2	Top	23.3	10.2	34.3	47.0	27.4	19.0
	Middle	8.6	Out	12.7	17.4	10.1	7.0
	Bottom	14.7	10.2	21.6	29.6	17.3	12.0
3	Top	15.8	8.4	24.4	38.2	22.8	12.4
	Middle	Out	Out	Out	11.6	7.0	Out
	Bottom	15.8	8.4	24.5	26.6	15.8	12.4
4	Top	10.5	5.8	16.7	26.1	13.5	9.5
	Middle	Out	Out	Out	10.0	Out	Out
	Bottom	10.5	5.8	16.7	16.1	13.5	9.5
5	Top	7.6	Out	11.5	14.7	10.0	7.3
	Bottom	7.6	Out	11.6	14.8	10.0	7.3
6	Top	5.5	Out	7.6	10.1	7.1	5.1
	Bottom	5.5	Out	7.6	10.0	7.1	5.1
7	Top	Out	Out	Out	5.4	Out	Out
	Bottom	Out	Out	Out	5.4	Out	Out
8	Top	Out	Out	Out	Out	Out	Out
	Bottom	Out	Out	Out	Out	Out	Out
	Total Duties	180.6	73.3	278.6	398.6	223.8	153.2

* For notation of guy positions see Fig. 6.

enced by the size of insulator units available. For example, it would not be economical to use insulator units where they carry only a small part of their rated duty.

It was, therefore, assumed inadvisable to insulate a guy where the total duty (with 150 kv. on the antenna) was less than 10 kv. or to increase the number of insulators in a guy when the total guy duty was increased less than 10 kv. In working out such an arrangement it was found that three insulators per guy was the largest number desirable. Four or five insulators per guy had very little more total duty than three insulators.

The results of an arrangement worked out on this basis

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are given in Table II and Fig. 15. No measurements were made with the mast insulated since at the time of the measurements it was not considered worth while to insulate the masts.

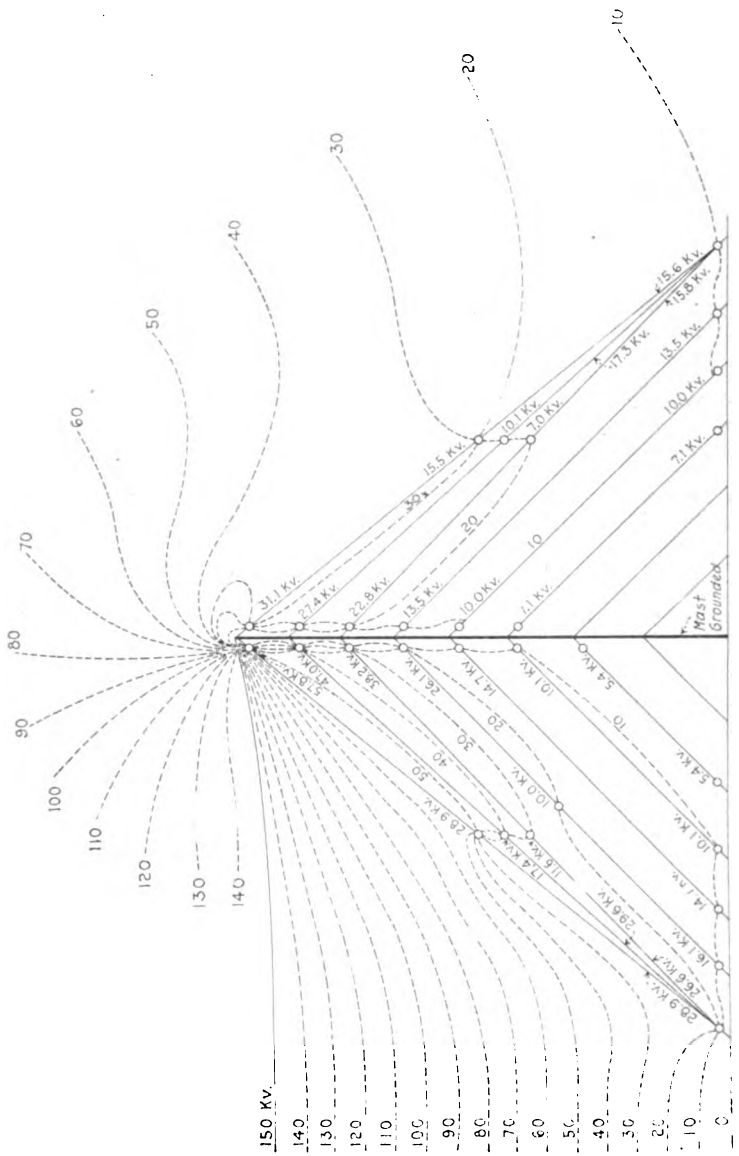

Figure 15—Insulator Duties and Electrostatic Field with 10 Kv. Minimum Guy Duty. Guy Positions NCS—Mast Grounded

The electrostatic field in Fig. 15 is very similar to that shown in Fig. 13, but it will be noted that a more economi-

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cal insulation arrangement has been effected. With only two additional insulators the total insulator duties for the entire mast have been increased from 1054 kv. (Fig. 13) to 1215 kv. (Fig. 15), a change of 18 per cent. In spite of this large increase it is seen that a still greater increase could have been obtained if the middle insulators of Fig. 15 had been arranged in a horizontal row instead of in a vertical row.

Conclusion

The final decision as to the value of mast and guy insulation will depend upon an economic study of the costs. The change in effective height caused by using a particular insulation arrangement can be measured on a model such as that described above and used to determine the saving in power effected. Balancing this saving against the cost of the insulators will show the value of the insulation. The various factors involved differ so much with the station and its location that such a balance for the example given above would be of interest only and of little practical value.

Acknowledgments

The writer is indebted to Dr. Harris J. Ryan of Stanford University for his many helpful suggestions in connection with this investigation and to Mr. C. H. Suydam for his assistance in making the measurements.

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DISCUSSION ON RADIATION RESISTANCE OF A VERTICAL ANTENNA —(LEVIN AND YOUNG)

Stuart Ballantine: With reference to the computation of radiation resistance at the base of a vertical antenna with excitation at unloaded harmonics, I should like to observe that these special values are very simply and accurately derived from the general formulae given in my paper (Proceedings, **12**, 823, 1924), particularly (18) in conjunction with the asymptotic developments (20) and (21) of the special functions $S_1(x)$ and $S_2(x)$ there employed. For odd harmonics $n=1, 3, 5, \ldots\ldots$, all the terms of (18) drop out except one so that the expression of the resistance at the antenna base is in this case reduced to the following simple form:

$$R_0=15\ S_1(2\pi n) \text{ ohms } \textit{(odd harmonics)} \qquad (1)$$

The functions $S_1(x)$ is related to the cosine-integral function, and has the following asymptotic development, useful for large values of n:

$$S_1(x)\sim \log x+.5772-\frac{1}{x^2}+\frac{6}{x^4}-\ldots\ldots\ldots (2)$$

For values of n above the first we may with considerable accuracy disregard the series, so that:

$$R_0=15(\log n+2.416) \text{ approximately} \qquad (3)$$

In view of the simplicity and accuracy of this formula, already published, I do not quite see why Messrs. Levin and Young were willing to undertake the somewhat tedious work connected with the derivation of their approximate formula. It may be of interest to compare the results given by the authors in their Table I with the values computed from (3); this tabulation follows:

Odd Harmonic (n)	Levin & Young	Comp. from (3)
1	37	36.2
3	52	52.7
5	60	60.2
7	65	65.4
9	68	69.2
11	71	72.2
13	73	74.9
15	75	76.9
17	77	77.7

*Received by the Editor October 11, 1926.

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For excitation at even harmonic frequencies the current at the base of the antenna is nominally zero and the values of radiation resistance which are frequently of the most interest in these circumstances are those corresponding to the point where the current amplitude is maximum. These values were previously designated as *R (loop)* and in the present case are given by:

$$R\ (loop)=60[S_1(\pi n)-\frac{1}{4}S_1(2\pi n)],\ \text{ohms} \qquad 4$$

where n—*2, 4, 6, ...* Now in view of (2) this may be written:

$$R\ (loop)\sim 45[\log n+1.491+\frac{5}{4}\ \frac{1}{(\pi n)^2}-\ldots\ldots\ldots] \quad (5)$$

or, with excellent approximation,

$$R\ (loop)\sim 45[\log n+1.491].\ \textit{even harmonics} \qquad (6)$$

Messrs. Levin and Young do not furnish numerical values for the even harmonics; the following computations from (6) may therefore be of interest.

Even Harmonic (n)	2	4	6	8	10	12	14
R (at current loop)	99.5 o	129	149	161	171	179	186

These values, together with those previously published, furnish us with the comprehensive picture of the variation of loop radiation resistance over the whole range of frequencies shown in Fig. 1. The loading necessary to produce resonance at the various frequencies is assumed to be inserted at the base, and the lower end of the vertical antenna is connected to a perfectly conducting earth. Due to the concentration of load at the base the maxima and minima of resistance do not coincide exactly with the unloaded harmonic frequencies. The divergence is greatest at low frequencies, decreasing as the frequency increases as would be expected, until the envelopes of the extermals are the dotted curves given by the equations (3) and (6).

As I have previously suggested (PROCEEDINGS, **13**, 255, 1925), when the antenna is erected over poorly conducting earth it is of some advantage to operate at an even harmonic, thus avoiding a great deal of earth loss. For short distance signaling by means of the direct earth ray the second *(n=2)* is suitable, while if the fourth *(n=4)* is employed the energy is radiated at an elevated angle of about 50 degrees and may cover greater distances by upper-atmospheric reflection. Some very excellent results with

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this mode of excitation have been reported to me in a private communication by B. A. Ostroumov, of the Radio Laboratory, Nijni-Nowgorod, Russia. In these tests, using a 15 k. w. Bontch-Brojewitch tube to supply a vertical antenna of 78 meters (or 95 meters) height at Moscow, Mr.

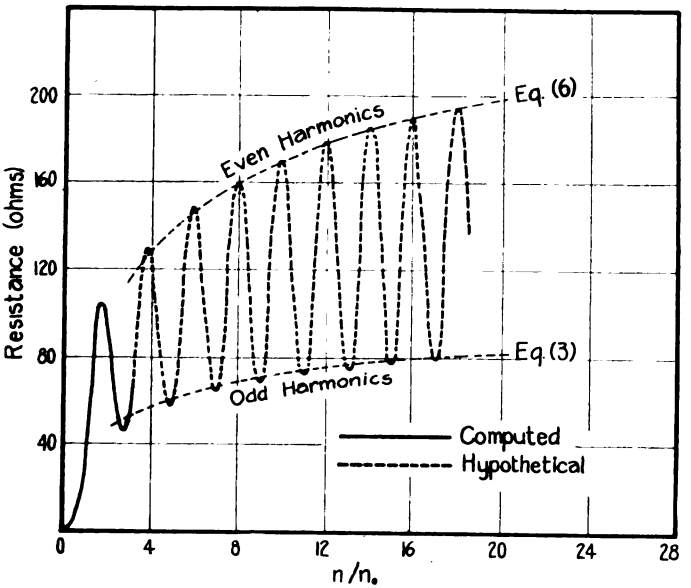

Figure 1—Radiation Resistance (at Current Loop) of Vertical Antenna over Perfect Earth and Loaded at Base as a Junction of Operating Frequency, Fundamental Frequency.

W. W. Tatarinov used, according to my earlier prescription, a wavelength of about 83 meters. Signals of considerable strength were reported at Porto-Rico and India.

In this note the term *harmonic* is employed in a strictly mathematical sense in which the fundamental is the first harmonic. This seems the best way to avoid the confusion which is current regarding the use of these terms. The term *harmonic* was originally borrowed from mathematics, whereas the physicist had been accustomed to referring to a *fundamental* and *overtones*. The *second harmonic* and *first overtone* are therefore synonymous, and so forth.

Some reference might have been made in this paper to the early work of Balth. van der Pol, Jr. on this subject.

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DISCUSSION ON THE OUTPUT OF CHACTERISTICS OF AMPLIFIER TUBES.—(WARNER AND LOUGHREN)

D. F. Whiting: I wish to take the opportunity that this occasion affords of emphasizing the importance to the radio engineer of the general subject with which the paper by Messrs. Warner and Loughren deals.

Mr. Warner has already pointed out to you that the development and use of improved loud speakers, which introduce less distortion of themselves and thereby allow the effects of tube overloading to be observed, has increased the need of amplifiers of considerably greater power.

To illustrate the similar effect of improvements in transformer design, let me say that puzzled radio-enthusiasts have often told me that, after replacing their old and supposedly inefficient audio-frequency transformers by those of improved design which are now appearing upon the market, they were must disappointed in the result of the change upon the quality of reproduction from their receiving sets. Upon questioning them regarding the type of tubes used and the conditions of use, such as grid and plate voltages, type of loud speakers used and volume of sound, it was at once apparent that they were badly overloading their amplifiers, which resulted in the poor quality that they had observed. The transformers which they had previously used cut off the low frequency where most of the energy of the original speech and music is concentrated, allowing the higher frequencies, to pass without serious overloading of the tubes. When transformers which were efficient at the low frequencies were substituted, the high energy of the low frequencies passed through the transformers to the tubes, overloaded them and produced the objectionable effect.

The obvious answer is to use not only a good loud speaker and good audio-frequency transformers, but also an amplifier possessing adequate power capabilities, the lat-

*Received by the Editor December 15, 1926.

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ter requirement being the subject of the paper under discussion.

Since the procedure for determining the optimum conditions and the maximum power output for a vacuum tube according to the method outlined in the paper involves the use of curves which are not easily obtainable by the average radio enthusiast, it might not be inappropriate to indicate, a few simple relations which will serve as a guide to the optimum conditions. These relations are partly theoretical and partly empirical; and, although absolute exactness is not claimed for them because of many minor effects which must be neglected, they are sufficiently accurate for most practical purposes, and the results obtained through their use have been found to agree fairly well with the results of careful tests.

Neglecting the case in which the power is limited by the energy radiation from the plate, a condition which is not usual in receiving equipment, and assuming that the filament is capable of emitting a copious supply of electrons, the grid bias E_c may be obtained from the expression

$$E_c = \frac{E_b}{2\mu}$$

where μ is the amplification constant and E_b is the potential at the plate of the tube, the negative filament terminal being considered the zero of potential. In case the amplification constant of the tube is unknown, a good value for E_c can be found by getting an indication of the plate current with E_c—*0* and E_b, normal and then increasing E_c until the plate current is reduced to about one-third of this value.

When using the proper value of grid bias, as determined above, and when operating into a resistance load equal to about twice the plate impedance, the maximum undistorted output power, $P_{max.}$, is approximately as follows:

$$P_{max.} = \frac{E_b^2}{36 r_p}$$

where r_p is the plate impedance at the plate and grid potentials chosen. In case the plate impedance is not known, $P_{max.}$ can be approximated in this way:

$$P_{max.} = \frac{E_b I_b}{10}$$

when I_b is the plate current which flows with a plate potential of E_b and a grid bias of E_c.

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I am not certain just how closely these relations will be found to agree with the method outlined in the paper, and it is not my purpose to provoke any controversy over their relative merits. Conditions are so affected by the criteria assumed for full load that exact comparison may be difficult. I believe however that these relations will be found useful even to those in possession of complete tube data; and I state them here to add to the discussion as an expression of this important subject.

A. V. Loughren: Mr. Whiting's quasi-empirical rules are of considerable interest and are quite useful when the operating conditions are those for which any particular rule was developed. Unfortunately however the introduction of an apparently minor change in the circuit may sometimes be sufficient to invalidate the rule. The authors have employed such rules to some extent in practice, but feel that their proper sphere is the checking roughly of a design quantity obtained from the more rigorous methods set forth in the body of the paper.

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DISCUSSION ON VACUUM TUBE NOMENCLATURE. (CHAFFEE)

Geo. R. Metcalfe: The system of nomenclature proposed by Prof. Chaffee has undoubted advantages for both the writer and the readers of mathematical articles. These are very clearly set forth in his paper. When manuscripts of this character reach the printer and publisher, however, numerous disadvantages become apparent which are not generally appreciated by the layman. The writer does not wish by any means to deprecate the value of mathematical papers, as many of the most fundamental contributions to engineering literature are couched in highly mathematical language. Unfortunately such papers are very expensive to print, consume an extraordinary amount of time in their mechanical preparation, and generally, when printed, appeal to a very small fraction of the membership of the societies by which they are published.

As an example of the expense of publishing an article using a nomenclature similar to that suggested in the paper under discussion; a forty-page paper of this character, recently published, cost for composition alone, $440.00 and for its complete publication about $800.00. If this paper was read and appreciated by twenty members of the society (and this is probably a liberal estimate) it cost the society forty dollars apiece to place the paper in the hands of these twenty members. In general, it may be safely estimated that a highly mathematical paper costs about five times as much as the same length of plain text, and in view of its very high cost and the relatively small number of readers, it is always a question as to how much money a society is justified in spending on a mathematical article. It certainly should be a paper of outstanding merit to warrant its publication at all.

From the foregoing it will be evident that in adopting a nomenclature the practical requirements of the printer should be given at least equal consideration with the convenience of the reader, in order not to stifle the production of technical literature by prohibitive costs. From this point

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of view the proposed nomenclature is unfortunate, because the use of one or two dashes above or below the letters and of caret signs and tildes over the letters doubles and triples the work of the compositor. A complicated equation with these various appendages attached to certain letters may, involve two or three hour's work and at the current rate of $3.50 per hour may bring the cost of five or six lines of type up to approximately $10.00. Such an equation consists of hundreds of different sized types all of which must be "justified" or fitted together so that the block may be lifted as a whole with each individual type tight in place.

From the standpoint of publication, any system of notation should be confined for the sake of economy to Roman, Italic and bold face capitals and small letters, and Italic superior and inferior English and Greek capitals and small letters and figures. These, with the necessary mathematical symbols, are sufficiently difficult to handle without introducing any further complications.

An incident occurred some years ago that illustrates the attitude of printers toward complicated mathematical nomenclature. A very important mathematical paper was presented before an engineering society and it contained quite a few special characters for which special matrixes had to be cut. In order to expediate the typesetting the manuscript was divided between two experienced compositors. The first one looked over his part and resigned his job at once. The second man worked until noon and then gave notice that he would leave the job that night. It took over six weeks to set that manuscript in type with less experienced compositors.

It is felt that anything that can be done to simplify the nomenclature is of distinct advantage to engineering, as the production of engineering literature will surely be curtailed by making it too expensive for the engineering societies to publish.

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DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

By John B. Brady

1,608,047—ALBERT H. TAYLOR and LEO C. YOUNG, Washington, D. C. Filed Jan. 7, 1925, issued Nov. 23, 1926. Assigned to Wired Radio, Inc.

RADIO SIGNALING APPARATUS wherein a radio transmitter may be located at a substantial distance from the transmitting antenna and the high frequency supplied to the antenna over a trunk line extending for a substantial distance.

1,608,048—ALBERT H. TAYLOR, Washington, D. C. Filed Jan. 23, 1926, issued Nov. 23, 1926. Assigned to Wired Radio, Inc.
Assigned to Wired Radio, Inc.

PIEZO-ELECTRIC CRYSTAL CONTROL SYSTEM for frequency control of transmitting stations wherein a protective circuit is provided for preventing undue strain upon the piezo-electric crystal element in the event that the current should rise through the piezo-electric crystal circuit below a predetermined safe value.

1,608,083—JAMES C. DALEY, Chicago, Ill. Filed Aug. 10, 1925, issued Nov. 23, 1926. Assigned to Jefferson Electric Mfg. Co.

MEANS FOR RECONDITIONING RADIO TUBES comprising an apparatus into which an electron tube may be inserted for successfully imposing high and low voltages upon the filament for predetermined periods of time for restoring the tube to operating condition.

1,608,085—H. A. DOUGLAS, Bronson, Michigan. Filed June 1, 1922, issued Nov. 23, 1926.

CONDENSER of variable capacity in which the disc-like plate may be moved within a cup shaped enclosure with respect to a stationary plate.

1,608,110—W. O. MEISSNER, Chicago, Ill. Filed Dec. 18, 1925, issued Nov. 23, 1926.

LOOP AERIAL in which a frame is provided with a plurality of apertures therearound and the conductors secured on the frame by members passing through the apertures.

1,608,146—A. M. TROGNER, Rakoma Park, Maryland. Filed Feb. 27, 1922, issued Nov. 23, 1926. Assigned to Wired Radio, Inc.

ELECTRICAL PROTECTIVE ARRANGEMENT for automatically disconnecting the plate potential in a high power electron tube transmitter should the antenna or other load accepting circuit fail to function.

1,608,272—P. E. GILLING, East Orange, N. J. Filed Mar. 19, 1925, issued Nov. 23, 1926. Assigned to Wireless Specialty Apparatus Co.

ELECTRICAL CONDENSER where a stack is secured under compression by means of U shaped clamps which slide over opposite ends of the condenser.

1,608,292—H. I. BECKER, Schenectady, N. Y. Filed Feb. 4, 1920, issued Nov. 23, 1926. Assigned to General Electric Co.

MEANS FOR PRODUCING ALTERNATING CURRENTS wherein a two element tube is provided with an external magnetic field control winding arranged in the oscillatory circuit.

1,608,146—A. M. TROGNER, Washington, D. C. Filed June 4, 1926, issued Nov. 23, 1926. Assigned to Wired Radio Inc., of New York.

MULTIPLE PIEZO-ELECTRIC HOLDER wherein a plurality of piezo-electric crystal elements of different frequency characteristics are mounted upon a rotatable frame which may be moved to selective positions for effectively positioning a desired piezo-electric crystal in the transmitter circuit.

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1,608,311—ALVARADO L. R. ELLIS, Swampscott, Mass. Filed Jan. 5, 1926, issued Nov. 23, 1926. Assigned to General Electric Co.

OSCILLATOR where a piezo-electric crystal element is connected to an electron tube circuit and the circuit tuned in response to a change in condition of the piezo-electric element for the control of a transmission circuit.

1,608,316—A. W. HULL, Schenectady, N. Y. Filed Jan. 29, 1920, issued Nov. 23, 1926. Assigned to General Electric Co.

METHOD OF AND MEANS FOR PRODUCING ALTERNATING CURRENTS which consists in producing a stream of electrons in substantially radial paths between an electron emitting cathode and a cooperating electrode and periodically substantially interrupting the electron stream between cathode and anode by means of a magnetic field which is in a direction substantially at a right angle to the paths of the electrons and which is produced by current carried by said electron stream.

1,608,317—AUGUSTUS C. HYDE, Perivale, England. Filed May 26, 1925, issued Nov. 23, 1926.

THERMIONIC VALVE having a cathode which is formed by coating a core of refractory metal by applying a solution of tungstic acid mixed with finely powdered thorium which is heated to at high temperature for forming the cathode.

1,608,429—C. E. RILEY, Charlestown, Indiana. Filed May 18, 1925, issued Nov. 23, 1926.

CRYSTAL DETECTOR HOLDER where a plurality of contact points in overlapping staggered relation are arranged to grip a crystal element therebetween.

1,608,472—J. J. AURYNGER, Brooklyn, N. Y. Filed Dec. 1, 1922, issued Nov. 23, 1926.

CONDENSER having a set of stator plates and two sets of rotor plates independently adjustable with reference to the stator plates.

1,608,504—G. W. HEATH, Newark, New Jersey. Filed April 15, 1924, issued Nov. 30, 1926.

ELECTRIC CONDENSER PLATE in which the surface of the plate is stippled with indentations for increasing the area of one plate with respect to the adjacent plate.

1,608,526—E. F. POTTER, Glencoe, Illinois. Filed Oct. 23, 1922, issued Nov. 30, 1926. Assigned to Kellogg Switchboard and Supply Company.

CONDENSER having a set of rotor and stator plates and a separate set of verier plates controlled by a concentrically positioned operating shaft.

1,608,535—P. SCHWERIN, New York, N. Y. Filed Dec. 30, 1921, issued Nov. 30, 1926. Assigned to Western Electric Co.

ELECTRIC DISCHARGE DEVICE in which the electrodes within the tube are supported on wire members flattened at selected points for positioning the electrodes in definite relation.

1,608,551—C. D. DEMAREST and R. K. BONELL, of Ridgewood and East Orange, New Jersey, respectively. Filed Dec. 30, 1925, issued Nov. 30, 1926. Assigned to American Telephone & Telegraph Co.

RADIO SIGNALING SYSTEM for linking line wire telephone systems with radio transmission and reception circuits with special circuit arrangements for indicating at a central station the particular secondary station which may be signaling the central station.

1,608,580—S. CABOT, Brookline, Mass. Filed Nov. 19, 1924, issued Nov. 30, 1926.

RADIO RECEIVING SYSTEM embodying a method of tuning a radio-receiving system of the radio-frequency amplifier type to the point of greatest sensitiveness for any wave length within the range of said system which consists, first, in simultaneously changing in opposite direction the impedance of the input and output circuits of a vacuum tube as wave length varies whereby any local oscillations that may be created in said system will be of substantially uniform intensity for any wave length within the range of said system, and, second, reducing the intensity of such local oscillations to as low a value as desired.

1,608,824—W. M. Sullivan, deceased; late of Newark, N. J. Filed Jan. 26, 1921, issued Nov. 30, 1926. Harriet Sullivan Payne, Administratrix.

SPARK GAP STABILIZER consisting of a member of conducting material disposed laterally of the path of rotation of the rotary electrode and connected to ground the rotary electrode being driven in synchronism with with the a-c. supply source of the set.

1,608,922—W. A. BOCKIUS, Wilmette, Ill. Filed June 6, 1925, issued Nov. 30, 1926. Assigned to H. F. Tideman, H. H. Whetter, and W. A. Bockius, of Chicago, Illinois.

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ELECTRICAL CONDENSER comprising a pair of sets of interleaved plates, pivoted to swing on different axes, and means for conjointly shifting the sets comprising a cam wheel connected to shift said sets.

1,608,969—H. W. WEBBE, Columbus, Ohio. Filed June 18, 1923, issued Nov. 30, 1926.

DIFFERENTIAL RADIOCONTROL wherein a transmitter may be modulated under selective control of tuning forks while tuning forks at the receiver may be similarly controlled for closing particular circuits.

1,608,974—A. S. BLATTERMAN, of Vail, Little Silver, N. J. Filed Jan. 29, 1921, issued Nov. 30, 1926. Assigned to Murad Radio Corp.

APPARATUS FOR CONTROLLING THE DIRECTIONAL RECEPTIO'' CHARACTERISTICS OF LOOP OR COIL ANTENNAE consisting of a plurality of sets of parallelly arranged conductors said conductors being of substantially equal length and arranged closely adjacent to a portion of the loop antenna to provide an electrical shield to said loop antenna. Means are provided for optionally connecting and disconnecting both sets of the parallel conductors with one another and ground, or only one set of said sets of parallel conductors with the receiver, the other set of conductors being in this instance entirely disconnected.

1,609,006—C. D. Tuska, Hartford, Conn. Filed Jan. 17, 1922, issued Nov. 30, 1926. Assigned to The C. D. Tuska Co.

VARIABLE CONDENSER wherein the rotatable shaft has a frictional bearing at one end consisting of a yielding strip anchored upon one of the insulated end plates of the condenser.

1,609,116—J. H. HAMMOND, Gloucester, Mass. Filed Aug. 17, 1922, issued Nov. 30, 1926.

METHOD AND SYSTEM FOR COMMUNICATION BY RADIANT ENERGY in which oscillations at the transmitter have periodic variations of relatively lower but supersonic frequency impressed thereon and varied in accordance with the signal at the receiver a special circuit is provided for transforming the energy into oscillations having a frequency intermediate between the frequency of the original oscillations and the frequency of the variations.

1,609,118—C. HARDY, New York, N. Y. Filed Dec. 10, 1925, issued Nov. 30, 1926. Assigned to Amsco Products, Inc.

ELECTRICAL CONDENSER MANUFACTURE where the plates of a condenser are keyed to a shaft by locking the plates within grooves formed in the shaft.

1,609,152—F. C. CARMEL, Pittsfield, Mass. Filed Feb. 9, 1925, issued Nov. 30, 1926. Assigned to A. H. Rice Co.

ANTENNA comprising a flat sheet of fabric including a plurality of braids laid parallel to each other and of insulating material and a conductor woven progressively forward in a zigzag manner across the sheet alternately between the strands of the sheet.

1,609,162—W. F. DIEHL, Jamaica, N. Y. Filed Dec. 19, 1924, issued Nov. 30, 1926.

RADIO RECEIVING APPARATUS of the superheterodyne type wherein there is a receiving circuit for the incoming electric wave, a second circuit adapted to be tuned to a frequency differing from the frequency of the first circuit, and a third circuit having a variable tuning element mechanically connected in such relation to a variable tuning element of the first circuit, as to maintain the third circuit, throughout the frequency range of the system, tuned to a frequency other than that at which the second circuit is designed to operate for the desired signal and differing from that of the first circuit by an amount equal to the frequency difference between the first and second circuits.

1,609,222—J. M. TAYLOR, Bridgeport, Pa. Filed Oct. 28, 1921, issued Nov. 30, 1926. Assigned to Diamond State Fibre Co.

INSULATED CONDUCTING ELEMENT for constructing condensers consisting of a body of metal foil with at least one sheet of parchmentized fibre applied thereto and extending through openings in said metallic foil.

1,609,398—C. E. BRIGHAM, of East Orange, N. J. Filed July 16, 1924, issued Dec. 7, 1926. Assigned to Brandes Laboratories, Inc.

VISUAL TESTING SYSTEM FOR ELECTROMAGNETIC SOUND REPRODUCERS by which the loud speakers may be tested both as to sensitivity and volume by observing their characteristics in an electrical circuit.

1,609,931—S. COHEN, New York, N. Y. Filed Oct. 30, 1923, issued Dec. 7, 1926.

ELECTRICAL CONDENSER of the fixed stack type where a baked coating of metallic oxide is formed directly upon the dielectric plates for forming the condenser.

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1,610,051—GUY HILL, Washington, D. C. Filed May 16, 1921, issued Dec. 7, 1926. Assigned to Wired Radio, Inc.

SUPPORTING PLATE FOR ELECTRICAL CONDENSERS in which a semi-circular metallic plate is embedded in an insulated plate and arranged to coact with movable plates for adjustment of capacity.

1,610,073—E. F. W. ALEXANDERSON, Schenectady, N. Y. Filed April 10, 1922, issued Dec. 7, 1926. Assigned to General Electric Co.

HIGH FREQUENCY SIGNALING SYSTEM in which two high frequency alternators are arranged to supply energy simultaneously to the same antenna circuit under control of separate modulator circuits without interference between the channels.

1,610,090—V. A. Hendrickson, Springdale, Conn. Filed Nov. 14, 1922, issued Dec. 7, 1926. Assigned one-third to Fred Berg and one-third to J. Edward Brown.

GENERATING AND CONTROLLING SYSTEM FOR RADIO TELEPHONY in which the anode circuit of the oscillator is provided with separate inductors which are coupled on opposite sides to the radiating inductance.

1,610,122—L. EDENBURG, Brooklyn, N. Y. Filed June 10, 1924, issued Dec. 7, 1926. Assigned to Dubilier Condenser & Radio Corp.

ADJUSTABLE CONDENSER comprising a helically wound strip which may be adjusted with respect to a correspondingly shaped strip for changing the capacity of the condenser.

1,609,366—FREDERICK A. KOLSTER, Washington, D. C. Filed Nov. 26, 1920, issued Dec. 7, 1926. Assigned to Federal Telegraph Company.

RADIO APPARATUS consisting of a closed circuit including an inductance and a capacity area where the capacity area constitutes a conductor which is a part of the inductance in the signaling circuit.

1,610,208—LESLIE R. McDONALD, Westmount, Quebec, Canada. Filed Oct. 31, 1925.

VACUUM TUBE VIBRATION DAMPER in the form of a rubber cup which snugly grips the bulb of the tube for damping vibration.

1,609,748—J. WEINBERGER, New York, N. Y. Filed Sept. 20, 1921, issued Dec. 7, 1926. Assigned to Radio Corp. of America.

RADIO TELEPHONE TRANSMITTING SYSTEM where alternating current above the limit of audibility is supplied to the plate filament circuit of an oscillator and the supply to the oscillator modulated in accordance with speech.

1,609,805—W. N. FANNING, Vallejo, Calif. Filed Jan. 23, 1922, issued Dec. 7, 1926.

RADIO RECEIVING SYSTEM wherein a relay is acted upon by both signal and static and the effect of static balanced out for the control of the relay by the signal.

1,610,258—A. E. CHAPMAN, London, England. Filed June 4, 1923, issued Dec. 14, 1926.

ELECTRICAL VARIABLE CONDENSER wherein two groups of plates are pivotally mounted spaced in position on a support with a third group of plates pivotally mounted adjacent the first two groups and all adjustable for controlling the tuning of an electrical circuit.

1,610,316—W. A. E. QUILTER, London, England. Filed Feb. 26, 1921, issued Dec. 14, 1926. Assigned to Radio Corp. of America.

THERMIONIC DEVICE wherein a grid electrode constructed in a plurality of metal parts, each part having apertures and being capable of movement for the adjustment of the size of the apertures for the control of electron discharge.

1,610,329—P. F. SHIVERS, Racine, Wis. Filed Oct. 19, 1922, issued Dec. 14, 1926. Assigned to Western Electric Co.

RADIO APPARATUS wherein the frequency of the transmitting station may be varied and a corresponding change effected at a receiving station for the secret transmission of signaling energy.

1,610,371—H. H. HAMMOND, JR., Gloucester, Mass. Filed (original) March 21, 1914; renewed July 24, 1923, issued Dec. 14, 1926.

GASEOUS DETECTOR OF RADIANT ENERGY AND METHOD OF CONTROL THEREOF, in which circuits are provided for the automatic interruption of the cathode and plate circuits of an electron tube system at a predetermined rate.

1,610,417—A. BOEDEKER and J. S. WENTWORTH, of Cincinnati and Norwood, Ohio, respectively. Filed March 7, 1925, issued Dec. 14, 1926. Assigned to Grandtone Radio Corp.

RADIO CONDENSER in which sets of plates are slidable with respect to each other in a frame of special construction for varying the capacity of the condenser system.

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1,610,425—E. L. CHAFFEE, Belmont, Mass. Filed Sept. 3, 1918, issued Dec. 14, 1926. Assigned to John Hays Hammond, Jr.

TRANSMISSION SYSTEM FOR RADIANT ENERGY wherein high frequency oscillations are produced, amplified, and then increased in frequency for impression upon a transmitting antenna circuit.

1,610,551—I. W. Eisenberg, Alhambra, Calif. Filed Mar. 28, 1923, issued Dec. 14, 1926.

RADIO EQUIPMENT wherein a variable condenser is arranged with a set of rotor plates and sets of alternately positioned stator plates. The stator plates are arranged in separate series, each insulated one from the other and separately connected to an electron tube system.

1,610,560—F. S. McCULLOUGH, Wilkinsburg, Pa. Filed Oct. 24, 1922, issued Dec. 14, 1926.

VACUUM TUBE FILAMENT CONSTRUCTION AND METHOD OF MANUFACTURING SAME for use in power tube devices. A pitted surface is provided for the cathode for increasing rigidity and strength for high power operation.

1,610,615—W. SCHAFFER, Berlin, Germany. Filed Oct. 28, 1924, issued Dec. 14, 1926. Assigned to Gesellschaft fur Drahtlose Telegraphie, M. B. H.

ARRANGEMENT FOR MAINTAINING ANODE VOLTAGE CONSTANT IN TUBE TRANSMITTERS, where a rectifier supplies the output circuit of an oscillator with capacity and loading devices traversed by the rectified current for maintaining the circuit anode potential steady.

1,610,704—S. D. PADDOCK, San Francisco, Calif. Filed Mar. 9, 1925, issued Dec. 14, 1926. Assigned to Paddock Engineering Corp.

MEANS FOR INCREASING THE CAPACITY OF RADIOAERIALS by providing metallic extension members upon a conductor comprising the antenna.

1,610,783—A. W. HULL, Schenectady, N. Y. Filed Dec. 18, 1920, issued Dec. 14, 1926. Assigned to General Electric Co.

ELECTRON DISCHARGE APPARATUS in which an electrode auxiliary to the normal three electrodes is employed and a positive potential applied for giving the device a negative resistance characteristic.

1,610,837—W. C. WHITE, Schenectady, N. Y. Filed Mar. 6, 1923, issued Dec. 14, 1926. Assigned to General Electric Co.

RECTIFIER CIRCUITS for obtaining uni-directional current from polyphase alternating current wherein individual rectifiers are provided in the separate phases and the effect of the rectifier integrated for the production of rectified current.

1,610,875—E. MAYER, Berlin, Germany. Filed Dec. 5, 1922, issued Dec. 14, 1926. Assigned to Gesellschaft fur Drahtlose Telegraphie.

ARC LAMP GENERATOR FOR PRODUCING AND AMPLIFYING ELECTRICAL OSCILLATIONS where a grid electrode is provided for controlling the passage of current between the anode and cathode.

1,610,919—E. BERLOW, East Orange, N. J. Filed Mar. 2, 1925, issued Dec. 14, 1926.

RADIO APPARATUS in which a hollow base is provided for receiving apparatus within which all of the wiring for the equipment within the received may be enclosed.

1,610,980—S. Silberman, Porz-on the Rhine, Germany. Filed July 18, 1925, issued Dec. 14, 1926.

ELECTRICAL CONDENSER built up in the form of a cable having a plurality of metallic sections each insulated one from another.

1,611,848—E. H. ARMSTRONG, New York, N. Y. Filed Dec. 13, (original) 1913; divided application filed May 8, 1926, issued Dec. 21, 1926. Assigned to Westinghouse Electric & Mfg. Co.

RADIO RECEIVING SYSTEM FOR CONTINUOUS WAVES consisting of a resonant receiving circuit upon which received waves are impressed, the resonant receiving circuit being coupled with a resonant detector circuit by which oscillations in the detector circuit react for effecting an audible signal in the receiving circuit.

1,611,990—C. E. BONINE, of Melrose Park, Pa. Filed May 28, 1924, issued Dec. 28, 1926.

ELECTRICAL CONDENSER in which a plurality of folds are provided for housing the plates of a condenser with an adjustable pressure device for regulating the special relation of the folds.

1,611,101—H. J. J. M. De R. DE BELLESCIZE, Paris, France. Filed August 29, 1921, issued Dec. 28, 1926.

RADIO SIGNALING SYSTEM for the transmission and reception of telegraph signals by a system of impulses which will make deciphering by a receiver not equipped with the special apparatus required for this system extremely difficult.

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1,611,182—WILLIAM H. FRASSE, Newark, N. J. Filed Dec. 4, 1924, issued Dec. 21, 1926.

RESISTANCE COUPLING which consists of a plurality of spaced terminals with resistance elements and an interposed condenser in series mounted in position with respect to the terminals for providing a coupling unit for radio receiving circuits.

1,611,183—WILLIAM H. FRASSE, Newark, N. J. Filed Dec. 4, 1924, issued Dec. 21, 1926.

GRID LEAK AND CONDENSER MOUNTING wherein a pair of spring arms are arranged to engage a condenser and a grid leak connecting the elements in shunt relationship.

1,611,215—A. P. McARTHUR, Bogoto, N. J. Filed Jan. 4, 1926, issued Dec. 21, 1926.

RADIO RECEIVING APPARATUS in which radio receiving apparatus is mounted upon a panel and arranged to be readily removed to a position within a cabinet or removed from the cabinet, at the same time establishing electrical connection with apparatus connected within the cabinet.

1,611,224—HARRY NYQUIST, Jackson Heights, N. Y. Filed Dec. 19, 1923, issued Dec. 21, 1926. Assigned to American Telephone and Telegraph Company.

METHOD AND APPARATUS FOR MEASURING FREQUENCY which consists in combining a given electro-motive force with an electro-motive force of slightly different standard frequency to produce a current whose frequency is the difference of the two electro-motive force frequencies. The magnitude of this difference may then be measured.

1,611,264—L. G. BURKWEST, Chicago, Ill. Filed Aug. 14, 1924, issued Dec. 21, 1926.

ADJUSTABLE CONDENSER in which a flexible metallic plate may be pressed in varying degrees with respect to a stationary plate for changing the effective value of the condenser.

1,611,313—W. LATRAVERSE, Montreal Canada. Filed Dec. 24, 1925, issued Dec. 21, 1926.

CONDENSER in which a plurality of variable condensers may be simultaneously operated through a system of gears with balancing plates for each condenser which may be adjusted with respect to each of the variable condensers.

1,612,134—C. V. LOGWOOD, San Francisco, Calif. Filed March 10, 1921, issued Dec. 28, 1926.

RADIO RECEIVING SYSTEM including an electron tube system having coupled grid and plate circuits with a mechanical interrupter arranged in the circuits in the form of a chopper.

1,612,259—W. R. BULLIMORE, Highbury Grove, London, England. Filed July 21, 1925, issued Dec. 28, 1926.

THERMIONIC VALVE wherein the capacity between the connecting pins of an electron tube is reduced by providing air dielectric between the pins.

1,612,284—J. H. HAMMOND, JR., Gloucester, Mass. Filed Oct. 14, 1922, issued Dec. 28, 1926.

SECRET RADIANT TELEPHONE where the transmitting station emits energy of varying character and the receiving station is provided with circuits which respond to the transmitted energy of varying character for piecing together the transmitted signals regardless of the disassociated arrangement of the signals which would be received upon any normal receiver.

1,612,285—J. H. HAMMOND, JR., Gloucester, Mass. Filed Oct. 31, 1922, issued Dec. 28, 1926.

SYSTEM OF SECRET RADIANTTELEPHONY where a transmitting station emits energy over a variable range of frequencies and means are provided in the receiving station for integrating the effects of the received energy upon an observation circuit.

1,612,414—W. C. BRUMDER, Milwaukee, Wis. Filed July 31, 1922, issued Dec. 28, 1926.

ADJUSTABLE CONDENSER including a pair of adjustable drums which may be actuated for varying the position of one condenser plate with respect to another condenser plate.

1,612,427—H. FLURSCHEIM, Paris, France. Filed Nov. 13, 1923, issued Dec. 28, 1926.

RADIO WARNING SYSTEM FOR USE ON VEHICLES consisting of a receiving installation for movable vehicles which will be actuated upon approach to a signaling station which might be located upon another movable vehicle for warning the occupant of the first vehicle of his approaching a position of danger.

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1,612,440—F. L. HUNTER, JR., Towaco, New Jersey. Filed Oct. 29, 1924, issued Dec. 28, 1926: Assigned to DeForest Radio Company.

ELECTRON DISCHARGE DEVICE where an insulated member is provided for supporting the electrodes within a tube, the insulating member being resiliently mounted within the tube.

1,612,448—M. LATOUR, Paris, France. Filed May 3, 1924, issued Dec. 28, 1926. Assigned to Latour Corp.

ELECTRICAL SIGNALING SYSTEM AND METHOD where a group of signaling stations are arranged to receive on a certain wave length and transmit on another wave length in cooperation with another group of stations with selective means for arranging the particular stations which are to be operated.

1,612,473—E. G. SHALKHAUSER, Peoria, Ill. Filed Aug. 27, 1926, issued Dec. 28, 1926.

VARIABLE CONDENSER in the form of a rotatable member which is screw threaded into a stator member in varying amounts for changing the capacity of the condenser.

1,612,476—W. K. THOMAS, Crafton, Pa. Filed Sept. 12, 1922, issued Dec 28, 1926. Assigned to Samuel A. Pickering.

RADIO RECEIVING SET in which a specific circuit arrangement is provided for impressing a positive potential on the grids of the tubes in the receiving circuit.

1,612,594—H. DE F. MADDEN, Newark, New Jersey. Filed Sept. 29, 1922, issued Dec. 28, 1926. Assigned to Westinghouse Electric & Mfg. Co.

MOUNT FOR ELECTRON EMISSION DEVICE in which a plate electrode is provided with side extensions which permit supports to be connected thereto where the supports extend in a vertical direction within the tube

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED JANUARY 5, 1927

Transferred to Fellow Grade

Pennsylvania,	E. Pittsburgh, Westinghouse Elec. & Mfg. Co.	Conrad, Frank

Elected to Fellow Grade

Germany,	Berlin-Lankwitz, 1 Luisenstrasse	Wagner, Dr. K. W.

Transferred to Member Grade

Massachusetts,	Arlington Heights, 72 Oakland Ave.	Lamson, H. W.
	Cambridge, 5 Craigie Circle, Suite 6	McElroy, Paul K.
New Jersey,	Englewood, 279 Audubon Road	Gillett, Glenn D.
	Roselle, 276 East 3rd Ave.	Hanson, O. B.
New York,	Brooklyn, 144 Underhill Ave.	Drew, C. E.
	Brooklyn, 274 Linden Blvd.	Pease, R. M.
	Rochester, c-o Stromberg-Carlson Te. Mfg. Co.	Graham, V. M.
Ohio,	Cleveland Heights, 3223 Ormond Road	David, B. W.

Elected to Member Grade

Dist. Columbia,	Washington, 1314 Belmont St. N. W.	Heron, L. M.
	Washington, 405 Ninth St. N. W.	Loomis, M. T.
Illinois,	Chicago, Armour Inst. of Tech.	Freeman, E. H.
New Jersey,	Plainfield, 730 Park Ave.	Barbulesco, C. D.
	So. Orange, 257 So. Ridgewood Road	Rodman, I. P.
New York,	Buffalo, 15 Wilbury Place	Million, J. W.
	New York City, c-o E. H. Robinson, 79 Madison Ave.	Moore, E. J. T.
Pennsylvania,	Scranton, 820 Wheeler Ave.	Graf, J. J.
England,	Hampshire, Farnborough, Victoria Road	Barber, W. E.
Sudan,	Kartoum, General Post Office	Edgeworth, K. E.

Associates Elected

California,	Hollywood, 1316 N. Sierra Bonita	Rogers, M. E.
	Hollywood, 1242 Gordan St.	Wilson, G. L.
	Los Angeles, 3338 W. 27th St.	Comyns, W. L.
	Los Angeles, 1922 Hooper Ave.	Pfleger, R. A.
	Pasadena, 2157 Dolores St.	King, A. P.
	Pasadena, 614 N. Holliston Ave.	Logan, M. A.
	San Francisco, 525 Market St.	Danforth, R. S.
	Sherman, 839 Elwood St.	Smith, W. H.
Colorado,	Denver, 6100 Montview Blvd.	Owen, R. H.
Connecticut,	Bridgeport, Bridgeport Herald	C. B. Kingston,
Dist. Columbia,	Washington, 6631 Harlan Place, N. W.	Waldo, G. V.
	Washington, 5503-33rd St. N. W.	Wheeler, H. A.
Illinois,	Berwyn, 3111 Grove Ave.	Wilkinson, L. G.
	Chicago, 6318 Kenwood Ave.	Andres, F. J.
	Chicago, 207 City Hall	Bronson, G. E.
	Chicago, 9551 Longwood Drive	Christ, J. F.
	Chicago, 2017 Grace St.	Cummings, G.
	Chicago, 716 Buena Ave.	Harries, H. M.
	Chicago, 5142 Race Ave.	Heasley, D. F.
	Chicago, 5480 Cornell Ave.	Hill, A. J.
	Chicago, 5005 W. Adams St.	Holmquist, C. T.
	Chicago, 1851 W. Adams St.	Howard, C. M.
	Chicago, 4607 Beacon St.	Jorgenson, H.
	Chicago, 2510 Lexington St.	Murphy, E. R.
	Chicago, 925 Wrightwood Ave.	Norene, R. E.
	Chicago, 24 S. Clinton St.	Pfeifer, P. S.
	Chicago, 6440 University Ave.	Temple, C. T.
	Chicago, 815 N. Pine Ave.	Trimble, L.
	Chicago, 3047 N. Robey St.	Weissmantel, C. H.
	Cicero, 2501 S. 61 Ave.	Zeithammel, A. J.
	Hinsdale, 37 N. Washington St.	Blair, W.
	Oak Park, 44 S. Maple Ave.	Blizat, C. R.
	Oak Park, 212 Marion St.	Phillips, L. F.
	Peoria, 507 Bigelow St.	Brewer, F. T
	Sycamore, 307 N. Main	Boynton, F. L.

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Indiana,	Valparaiso, 451 Greenwich St.	Piatt, F. W.
	Warsaw, 9 Times Bldg.	Franklin, T. U.
Iowa,	Council Bluffs, 2408 S. 12th St.	Kelly, E. L.
Kansas,	Manhattan, 1622 Houston St.	Whan, L. H.
Louisiana,	New Orleans, 844 Howard Ave.	Clemmons, W. A.
	New Orleans, 2211 Octavia St.	Green, W. H.
	New Orleans, 321 St. Charles St.	Nall, V.
	New Orleans, 6036 Patton St.	Nebe, H. G.
	New Orleans, 512 St. Peter St.	Pohl, J .A.
	New Orleans, 139 S. Rampart St.	Willoz, V. L.
Massachusetts,	Cambridge, M. I. T. Dormitories	Chinn, H. A.
	Cambridge, 984 Memorial Drive	Fahnestock, H., Jr.
	Cambridge, 34 Mass Ave.	Gebhardt, P. B.
	Winchester, 16 Wildwood St.	Allen, A.
Michigan,	Detroit, 1650 W. Gd. Blvd., Apt. 5,	Lyons, F. E.
	Detroit, 774 Casgrain Ave.	Waynich, A. H.
Minnesota,	St. Paul, 820 Gt. Nor. Bldg.	Rankine, J. C.
New Jersey,	Bayonne, 89 W. 26th St.	Dillenbeck, P. D.
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	Montclair, 54 Hawthorne Pl.	Hervey, J. P.
	Wenonah, 109 E. Jersey Ave.	Beeler, E.
New York,	Brooklyn, 918 E. Parkway	Brody, M. M.
	Brooklyn, 2016 Greene Ave.	Grade, F. E.
	Brooklyn, 1705-77 St.	Hamre, F.
	Brooklyn, 88 Livingston St.	Michelson, C. D.
	Brooklyn, 255 Hull St.	Redfern, F .F.
	Brooklyn, 1907 Homecrest Ave.	Reynolds, J. B.
	Brooklyn, 195 Ocean Ave.	Shestacovsky, J.
	Brooklyn, 4821-11th Ave.	White, P. S.
	New York, 41 Park Row	Allyn, R. S.
	New York, 463 West St.	Anderson, W. H.
	New York, 10th Ave. & 36th St.	Baukat, H. W.
	New York, 453 W. 143rd St.	Brouthers, C. A.
	New York, 168 Washingtotn St.	Fernald, P. R.
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	New York, 16 E. 40th St.	Henry, C. C.
	New York, 326 Broadway	Israel, J. J.
	New York, 233 Broadway	Melhuish, H. T.
	New York, 463 West St.	Mellor, W. N.
	New York, 1295 Madison Ave.	Purdy, F. R.
	New York, 136 Liberty St.	Schmetf, F. E.
	New York, 571 W. 130th St.	Turski, A.
	New York, U. S. S. Seattle, Care Postmaster	Whitaker, J. N.
	New York, 920 Prospect Ave.	Wright, H. M.
	Port Jefferson, 1207 Main St.	White, L. S.
	Schenectady, Radio Eng. Dept. G. E. Co.	Hultkrantz, J. G.
	Scotia, Route No. 2	Palmer, H. L.
	Syracuse, 412 S. Alvord St.	Blaich, J. R.
Ohio,	Cincinnati, Brunswick-Bulke-Collinder Co.	Albright, G. F.
	Cleveland, 8520 Carnegie Ave.	Calkins, W. G.
	Cleveland, 1101 Union Trust Bldg.	Covert, C. B.
	Cleveland, 2028 W. 99th St.	McCollister, D. R.
	Cleveland, 464 East 117th St.	Robertson, F. E.
Pennsylvania,	Altoona, 2629 Maple Ave.	Aughenbaugh, W. K.
	E. Pittsburgh, Westinghouse	Landon, V. D.
	Lansdowne, 79 S. Wycombe Ave.	Kimball, G. W.
	Norristown, 739 Noble St.	Sands, W. F.
	Philadelphia, 1912 Walnut St.	Kennedy, T. R.
	York, 434 N. Beaver St.	Gemmill, W. E.
Vermont,	Northfield, 11 Highland Ave.	Howes, D. E.
Virginia,	Arlington, Naval Radio Station	Freed, R. E.
W. Virginia,	Wheeling, 42 Poplar Ave.	Weimer, E. W.
Wisconsin,	Madison, 725 W. Dayton	Borden, F. G.
	Madison, 222 N. Orchard St.	Davis, R. J.
	Madison, Burgess Laboratories	Martin, T. J.
	Milwaukee, 747 Summitt Ave.	White, W. M.

Associates Outside the U. S.

Brazil,	Sao Paulo, Mackenzie College	Weeden, E. H.
Canada,	Montreal, 497 Phillips Square	Crawford, J.
	Sudbury	Jessop, F. R.
	Toronto, 1936 Queen St.	Emes, K. C.
	Toronto, Apt. 16, "The Oaks"	De Grave, C.
	Toronto, Prince George Hotel	McClain, G. W.
China,	Shanghai, Commercial Press	Yu, Chia Chang

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England,	Bournemouth, 1 Irving Rd., West Southbourne	Kilmister, G. W.
	London, 11 Brassie Ave., East Acton	Jervis, F. J.
	London, 27 Stanwick Road, West Kensington	Leathes, W. H. B. deM
	Natwich, Coppice Road, Fern Villa	Noden, J.
Mexico.	Mexico City, Gante No. 20	Domenzain, S.

Juniors Elected

Connecticut,	West Hartford, 111 South Highland Ave.	Wagner, R. P.
Georgia,	Decatur, 1127 Clairmont Ave.	Hamilton, J.
Illinois,	Chicago, 1407 Irving Park Blvd.	Imes, R. N.
Massachusetts,	Brookline, 62 Chestnut St.	Duffy, J. A.
New Jersey,	Clifton, 370 Clifton Ave.	Watts, I. B.
Pennsylvania,	Philadelphia, 2070 West 65th Ave.	Hutchby, G. E.
Texas,	Houston, 113 N. Palmer	Wheeler, H.
Wisconsin,	Madison, 2246 Talmadge St.	Zurian, P. D.

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 16 APRIL, 1927 Number 4

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GENERAL INFORMATION

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INSTITUTE ACTIVITIES

March Meeting of Board of Direction

At the meeting of the Board of Direction, held at Institute Headquarters on March 2, 1927, the following were present: Dr. Ralph Bown, President; Frank Conrad, Vice-President; Dr. A. N. Goldsmith, Secretary; W. F. Hubley, Treasurer; Dr. J. H. Dellinger, Junior Past President; Donald McNicol, Junior Past President; Melville Eastham, L. A. Hazeltine, J. V. L. Hogan, R. A. Heising, R. H. Manson, R. H. Marriott and J. M. Clayton, Assistant Secretary.

The following were transferred to the grade of Member: Walter Van Nostrand, Jr., Ralph M. Heintz, Charles R. Rowe, Earl S. Fletcher and Charles H. Stewart.

The Board approved the election of the following to the Member grade: Charles C. Shumard, F. L. Pendergrass, W. R. McCanne, I. G. Maloff, E. S. Rogers, Dr. G. Leithauser, L. E. B. Everett and Albert Hall.

Eighty-five Associate and nineteen Junior members were elected.

A petition, from Institute members residing in Detroit, for the formation of a Detroit Section of the Institute was presented and the formation of the Detroit Section was approved.

New York Meeting of the Institute

The meeting of the Institute was held on March 2, 1927 in the Engineering Societies Building, 33 West 39th street. A paper by Dr. A. Meissner on, "Piezo-Electric Crystals at Radio Frequencies" was read by Dr. K. S. Van Dyke of Wesleyan University.

The paper was discussed by Dr. J. H. Dellinger, Dr. K. S. Van Dyke and others. The attendance at this meeting was over two hundred.

Meeting of Committee on Admissions

At the meeting of the Committee on Admissions held in the offices of the Institute on March 2nd, the following were present: Frank Conrad, Chairman; Messers. Hazeltine, Heising, Hull and Kroger. The Committee approved

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nine applications for election to the grade of Member and four applications for transfer to this grade. The Committee failed to approve the applications of two Associates for transfer to Member grade, and two non-members for direct election to this grade of membership.

1927 Year Book

The 1927 Year Book probably will be mailed to the entire membership within the next month. This Year Book will contain the names and addresses of the members of the Institute together with much information of value to the members.

The list of membership will appear in geographical as well as alphabetical order.

Each member is requested to advise the office of the Institute of any incorrect listing of names or addresses.

Morris Liebmann Memorial Prize

For a number of years the donor of the fund from which the annual award of the Morris Liebmann Memorial Prize has been made possible has requested that his name be withheld.

At the Banquet of the Convention it was announced that the donor was Emil J. Simon, a Fellow of the Institute and a well-known radio engineer.

Application for Member and Fellow Grade

The work of passing upon applications for admission or transfer to the grade of Member or Fellow will be aided materially if the applicant, at the time of filing his application for admission or transfer, will write to his references giving them a copy of his training and experience record as it appears upon his application form.

In many cases the persons to whom the applicant refers are acquainted with the applicant but do not recall the details of his past training and experience and are therefore unable to give information of value to the Admissions Committee.

In general, applications for the grade of Junior or Associate can be accompanied with references from non-members of the Institute, when the applicant is not acquainted

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with five Associates, Members or Fellows of the Institute, if the names of five of the applicant's business associates or friends are given. These should be men engaged in engineering or scientific work, if possible.

News of the Sections

Rochester Section

The Rochester Section held a meeting on the roof of the Sangamo Hotel, Rochester, N. Y., on February 11, 1927. The meeting was addressed by George C. Furness who described in detail the work required in preparing and presenting the "Eveready Hour".

Through the courtesy of station WHEC, the program was broadcast.

The attendance at this meeting was two hundred.

Boston Section

A meeting of the Boston Section was held on February 25th in Cruft Laboratory, Harvard University. Glenn Browning presented a paper on "Inter-Stage Tuned Radio Frequency Transformers". In the discussion which followed, the following participated: Professor Pierce, Professor Field, Messrs. Murray, Lamson, Brown and others.

The attendance was seventy.

Philadelphia Section

On February 25, 1927, the Philadelphia Section held a meeting in the Bartol Laboratories, 131 N. 19th street, Philadelphia. Paul E. Ritter delivered a paper on, "A Talk on Telephotography". A general discussion followed.

At this meeting there were thirty-five members present.

The next meeting of the Philadelphia Section will be held on March 25, 1927 in the Bartol Laboratories.

Los Angeles Section

A meeting of the Los Angeles Section was held on February 21, 1927 in the Los Angeles Commercial Club. Talks were given by Messrs. Wallace and Heller on, "Short Wave

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Experimental Work" and, "Master Oscillators in Broadcast Transmission". The talks were discussed by Messrs. Heller, Nikirk, Leighton, Wallace and Waters.

The attendance was thirty-five.

Chicago Section

The Chicago Section held a meeting on February 25, 1927 at which Professor G. M. Wilcox presided. The meeting was held in the auditorium of the Western Society of Engineers. Two papers were presented, one by Professor E. M. Terry on, "Factors Affecting the Constancy of Quartz Piezo-Electric Oscillators", and the second by E. S. Andrews on "Electrolytic Condensers for "A" Current Filters".

The papers were discussed by Messrs. Arnold, Marco, Adair, Miller, Terry, Brack and others.

The next meeting of the Chicago Section will be held in Fullerton Hall, Art Institute, Chicago on March 18th. A. A. Oswald of the Bell Telephone Laboratories will deliver a paper on, "Trans-Oceanic Telephone Transmission".

Washington Section

A meeting of the Washington Section was held in Harvey's Restaurant, 11th and Pennsylvania Avenue, Washington, on March 9th. C. Francis Jenkins read a paper on, "Visual Radio". The presiding officer at this meeting was Captain Guy Hill.

Thirty-five members attended the meeting.

The next meeting of the Washington Section will be held on April 13th at which time Dr. L. P. Wheeler of the U. S. Naval Research Laboratory will deliver a paper on, "Establishment of Standard Radio Frequencies".

Detroit Section

A meeting of Institute Members residing in Detroit was held on February 18, 1927. This is the third of the preliminary organization meetings looking to the formation of the Detroit Section.

Thomas R. Clark presided and a talk was given by

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Earle D. Glatzel on, "Development of Power Line Carrier Systems". The attendance was thirty-five.

Since this meeting was held the Board of Direction has approved the petition submitted by the Detroit members, and the Detroit Section has been formed.

The next meeting will be held on March 18, 1927.

Proposed Section at Cleveland

For some time correspondence has been carried on between Institute Headquarters and members residing in and around Cleveland, Ohio, for the purpose of establishing a Cleveland Section of the Institute.

On February 25, 1927, an organization meeting was held in the Lecture Room of the Physics Building of the Case School of Applied Science in Cleveland. The meeting was called to order by B. W. David. Colonel L. R. Krumm, acting as the Institute's representative, delivered a talk on the Institute, Sections and benefits to be derived from Sections.

Officers of the temporary organization were then elected as follows: John R. Martin, Chairman; R. E. Farnham, Vice-Chairman; L. L. Dodds, Secretary-Treasurer. The following temporary Committee Chairmen were appointed: Committee on Meetings and Papers, D. Schregardas and Publicity Committee, B. W. David.

Following the business part of the meeting, D. E. Replogle of the Raytheon Manufacturing Company delivered a paper on, " 'B' Power Supply Devices".

A petition requesting recognition of the proposed Cleveland Section has been forwarded to Institute Headquarters.

Committee Work

Sectional Committee on Radio, A. E. S. C.

The following is a report of the activities of the Technical Committees of the Sectional Committee on Radio, A. E. S. C., for January 1927:

Committee on Transmitting and Receiving Sets and Installations

A number of changes in the personnel of this Committee have been made during the month. The membership of

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the Committee with their respective affiliations is as follows:

National Electric Manufacturers Association

C. J. Young (General Electric Company)
J. D. R. Freed (Freed-Eisemann Corporation)
W. H. Leathers (Graybar Electric Company)
G. Lewis (Ken-rad Corporation)

Pacific Radio Trades Association

E. M. Sargeant

Association of Electragists

A. L. Abbott

Bureau of Standards

J. H. Dellinger

Inter-Department Radio Advisory Committee

J. H. Dellinger

Institute of Radio Engineers

M. Eastham (General Radio Company)

American Institute of Electrical Engineers

W. R. G. Baker (General Electric Company)

National Fire Protection Association

H. B. Smith (Underwriters Laboratories)

Committee on Component Parts and Wiring

A meeting of this Committee was held at the American Electrical Railways Association Board Room on February 2nd, 1927. A Subcommittee was appointed to investigate variable condenser standardization. No standardization material was adopted for recommendation to the Sectional Committee, the Committee awaiting reports from the various Subcommittees which have been investigating various subjects for standardization. The next meeting of the Committee will be on March 2nd, 1927.

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Committee on Vacuum Tubes

The Chairman of this Committee has been active this month in an attempt to secure exact dimensional data on bases of vacuum tubes for possible standardization.

Committee on Electro-Acoustic Devices

Some material for consideration by the members of the Committee has been received from Dr. Hund of the Bureau of Standards. No meeting of the Committee was held during January.

Committee on Power Supply and Outside Plant

No action has been taken by this Committee during January, 1927.

Meeting of Subcommittee on Receiving Sets

A meeting of the Subcommittee on Receiving Sets, of the Standardization Committee of the Institute, was held in the Institute offices on March 3rd. Those present were J. H. Dellinger (Chairman), A. F. Van Dyck, T. A. Smith, V. M. Graham, E. Austin, L. A. Hazeltine, E. E. Hiler, G. C. Crom, J. V. L. Hogan and L. C. F. Horle.

The meeting was largely devoted to the perfecting of the descriptions of basic methods of testing three overall characteristics of receiving sets, for which a preliminary draft had been prepared at a previous meeting. These three overall characteristics are sensitivity, selectivity, and fidelity. The Subcommittee prepared definitions of these terms and outlined testing methods and standard ways of expressing the results. The apparatus required for the measurements is a source of radio-frequency voltage, of adjustable frequency, intensity, and modulation percentage, a standard artificial antenna (if the receiving set does not have a self-contained antenna), and instruments to measure audio-frequency power output. The results of the measurements are expressed in graphs.

The Subcommittee has not completed this work. Some of the features of the tests remain to be specified by further work of the Subcommittee.

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PIEZO-ELECTRIC CRYSTALS AT RADIO FREQUENCIES *

By
A. Meissner
(Telefunken Company, Berlin, Germany)

Piezo-electric crystals were first introduced into practice by Cady[1] for the control of high-frequency oscillations and for the production of oscillations used in transmitter control. In general, quartz plate elements, cut parallel to the optical axes and perpendicular to one of the electrical axes, are employed in practice as piezo-electric bodies. The plate element (Fig. 1) has three natural periods of mechanical vibration to correspond to its three dimensions. If the

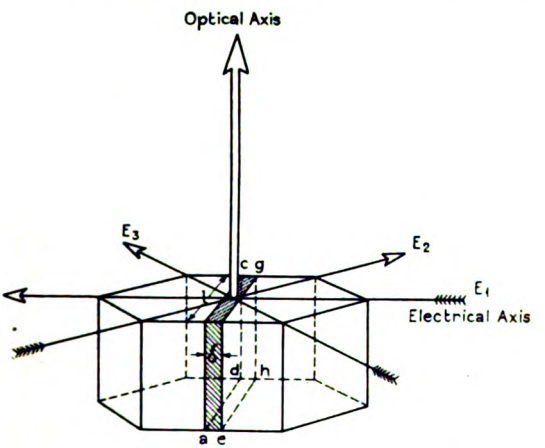

Figure 1—Orientation of Quartz Plate

plate element be excited by radio frequency (electrodes on faces *abcd* and *efgh*), two of the natural periods of mechanical vibration—to wit, the transverse, corresponding to the thickness δ, and the longitudinal, corresponding to the length l—will appear as natural electric oscillations[2].

*Translated from manuscript received October 11, 1926. Presented at the Institute of Radio Engineers Meeting, New York, N. Y., March 2, 1927.
[1]U. S. A. Patent No. 1,450,246 (Wave Control) and No. 172,583 (Generation of Oscillations).
[2]Literature: See Scheibe's Summary, Jahrbuch, Bd. 28, Page 15.

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Quartz crystals were first applied to wave control. A large number of circuits were devised for this purpose. A very interesting application is that of Mr. Giebe; namely, the utilization of the glow, such as the glow effect in a vacuum, arising from a high-frequency excited crystal.

In order to make the adjustment to any desired wave as simple as possible and the luminous effect at crystal resonance as sensitive as possible, a wave control was tried such as is shown in Fig. 2. *A* is a circuit excited at the controlled frequency.

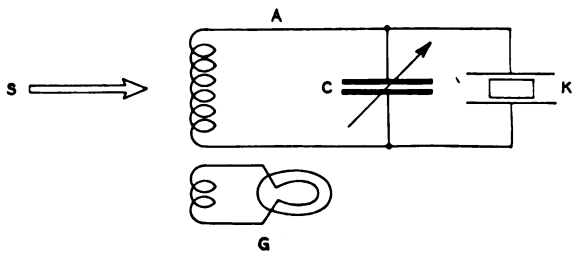

Figure 2—Wave Control by Means of an Incandescent Lamp.

The crystal lies parallel to the tuning condenser *C*. A small incandescent lamp *G* is coupled to this circuit. The lamp lights when the circuit *A* is roughly tuned to the transmitter frequency by the variable condenser *C*. If the wave of the transmitter to be controlled is exactly tuned to crystal resonance, the crystal draws so much energy from the tuned circuit that the lamp dims. The incandescent lamp is a much more positive light indicator for this purpose than the self-luminous crystal (in a vacuum). Besides, the adjustment to crystal resonance by means of the dimming of an incandescent lamp as an indicator is much sharper than by the method of adjusting to highest luminosity. The adjustment from darkness to luminosity is to a certain extent a method of minima. Quite astounding figures of precision of adjustment—within one part in 100,000 or more—are attainable with this method.[1] Such sharpness of tuning is not of paramount importance. It may be only a hindrance. For the furtherance of crystal technique, it was first necessary to establish a method for the study of crystals, an arrangement for positive indications of resonance maxima, a method which would enable us to reproduce the

[1]Literature: See Scheibe, "Jahrbuch," Bd. 28, beginning p. 21.

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resonance curve of the crystal. Fig. 3 shows a circuit of Dr. Heegner's for this purpose. An amplifier tube is used. The transmitter acts only by induction. If the crystal K is excited to resonance, an indication is obtained on the instrument J in the plate circuit. We get the resonance curve by varying the frequency of the transmitter and reading the current indications of J.

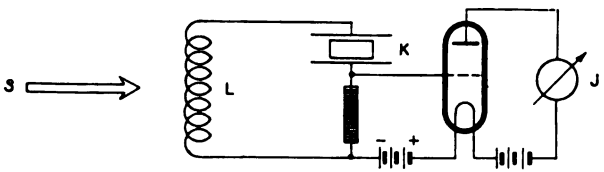

Figure 3—Wave Control by Means of a Vacuum Tube.

A still simpler arrangement resulted through the combination of a crystal with a detector; e.g., as in Fig. 4. The crystal transfers part of its oscillation energy to the detector. The transmitter S acts upon the coil L. If the frequency of S is varied J will indicate the current values of the resonance curve. At resonance the crystal transfers maximum energy to the detector.[1] This manner of reproducing the resonance curve offers us a method, free from objection, for obtaining the values of the damping of the crystal bodies, as well as a way for studying all mechanical and electrical effects on crystal oscillation. Curve 1 (Fig. 5) gives the resonance curve of the transverse oscillation; curve 2, of the longitudinal oscillation. The damping of the crystal for transverse vibration is $d=0.000\ 04$ $(\lambda=500\ m.)$; the damping for longitudinal oscillations, $d=0.00012$.

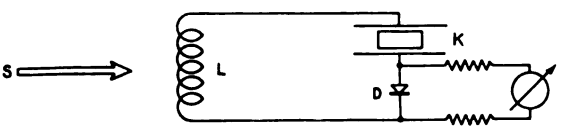

Figure 4—Wave Control by Means of a Detector.

[1]The somewhat small and disturbing steady deflection of the detector at wave lengths under 500 meters can be minimized in a simple way.

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Indeed, we have been able to find means to modify the resonance curves of crystals.

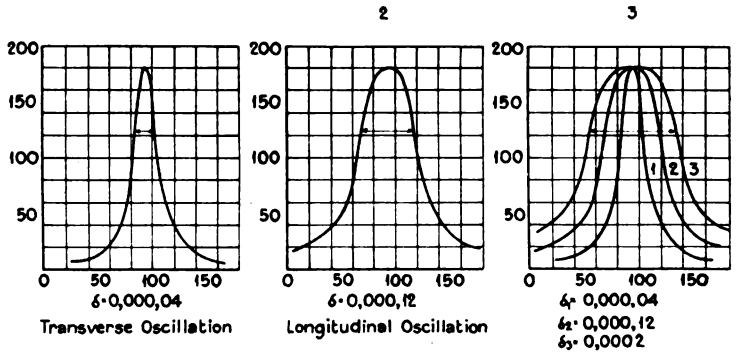

Figure 5—Damping of Crystals.

The damping can be changed either electrically or mechanically. Curve 3 (Fig. 5) shows the crystal of curve 1 damped by mechanical means. Curve 2 shows the same crystal damped electrically. The damping can be increased tenfold or twentyfold.

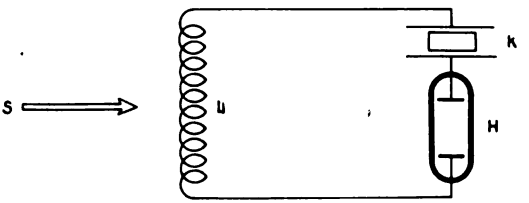

Figure 6—Wave Control by Means of a Glow Tube.

Other methods were found for proving and demonstrating crystal resonance. Fig. 6 shows one such method. In this method, a glow tube is in series with the crystal which consists of a single quartz plate element between two electrodes in an air medium.* The glow tube is separate from the crystal. In the figure, it is in series with the crystal. If the crystal is excited, its energy will be imparted to the glow tube. The glow tube brightens at the maximum of resonance.†

Following are the advantages of a wave control system of this kind:

*(Translator's Note: "in an air medium", as distinguished from some other medium, such as oil.)

†One must see that the capacity of the crystal mounting is not too great.

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1. As the crystal is a self-contained unit, it can be made for any, and every wave length.
2. The crystal can be damped to avoid excessive sharpness of resonance.
3. The crystal operates at its optimum excitation and its best utilization factor.
4. The resonance indication is given by large glow tubes with greater light emission.

Thus, the question of wave control seems to be satisfactorily solved technically. However, we must now ask whether, on the whole, the control of the transmitter in this way, by means of a wave standard, is a practical solution. All these methods of control are very complicated. First, one must juggle to and fro with the transmitter, until one finally hits on the usually excessively sharp resonance point of the crystal. Such a method is generally too difficult for the type of personnel that we must count on in engineering. The most convenient and logical solution is the following: To fix the frequency of the oscillator from the very beginning by means of a crystal, using the crystal as a control and as a calibration source for the transmitter, so that it will be impossible for the transmitter to oscillate at a frequency other than the normal crystal frequency. The crystal

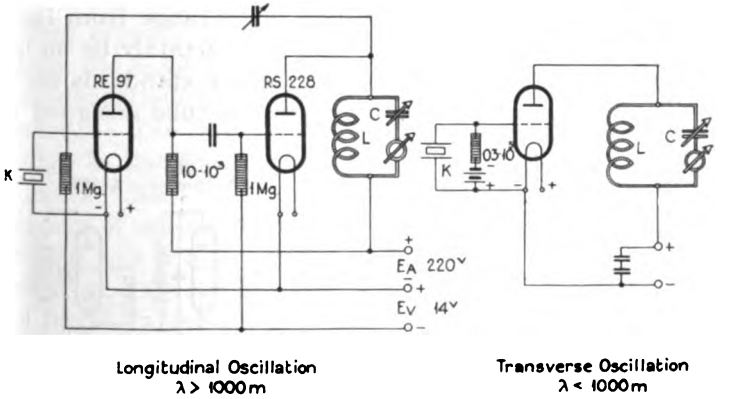

Figure 7—Crystal Oscillator.

must be used as an oscillator only. The crystal is placed together with a resistance and a choke in the grid circuit of an amplifier tube; the plate circuit is tuned. This circuit is only to be used for the transverse oscillation of the crystal. For this purpose, the limit of thickness of crystal for

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the shorter waves *(λ=100 m.)* is about 1 mm; and, for practical reasons, not over 1 cm. for the longer wave lengths *(λ=1,000m)*.

A harmonic of the 1 mm. crystal is used for shorter wave lengths; while the longitudinal oscillations of the crystal are of importance for wave lengths over 1000 m. The circuit for such oscillations is the two-tube outfit of Dr. Heegner (Fig. 7, left). The crystal is in the circuit of the first tube, the tuned circuit in the circuit of the second tube.

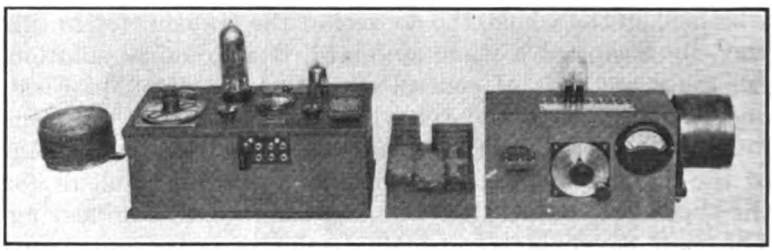

Figure 8—Small Test and Measuring Crystal Transmitters with Wave Range from 100 to 8000 Meters.

There is regenerative capacitive coupling between the tuned circuit and the crystal. The photographs (Fig. 8) show two small test and measurement transmitters of 1 to 3 ohms connected to 220 to 440 V.), with a wave range from 100 to over 8,000 m. In the future, there will certainly be no lack of these small transmitters as wave length standards for all testing purposes. The first unit has one tube designed for

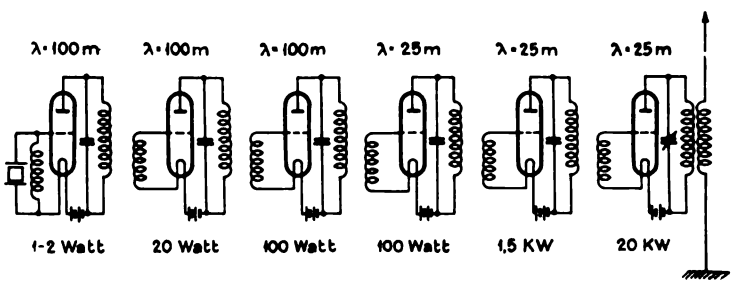

Figure 9—Crystal Excitation for a 25-Meter Transmitter.

100 to 1,000 m. The second unit has two tubes and is designed for wave lengths over 1,000 m. The crystal lies between two metal surfaces which are held in position by a hard rubber mounting. The crystal for the desired wave

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length is inserted in each of the two units at the left and right, respectively, of the containing box. The condenser is set to the calibrated value; the deflection of the ammeter in the tuned plate circuit serves as an indication that oscillations have set in. Crystal excitation is now the universal solution for purposes for which a very exact maintenance of transmitter frequency is required. This solution is, therefore, especially important for short wave lengths. Fig. 9 shows the arrangement for the two short wave transmitters at Nauen, for wave lengths from 25 to 40 m. 100 meters is the wave length of the crystal oscillation. The energy output is stepped up to 100 watts by two tubes, then quadrupled and passed through three more amplifiers. The final energy in the antenna is 10 kw. Crystal excitation is also of importance for broadcast transmitters.

A whole series of broadcast stations in America operate with crystals. Our setups for this purpose are identical with those shown in Fig. 9. The stages of amplification back of the crystal tube are 5 w., then 75 w., 500 w., and 3 kw. in the antenna.

An attempt was made to solve the present-day problem in present broadcasting of single-wave broadcasting by several stations utilizing crystal transmitters. The solution therefore consists in giving two separate transmitters controlling crystals ground to the same wave length; i.e., to exactly the same thickness. The studies made showed that a disproportionately high accuracy in the wave equality is necessary. The two transmitters must not vary from one another by more than 5-10 cycles; i.e., their frequency must be exact within 10^{-3} to 10^{-4} per cent.* Naturally, such accuracy of agreement can not be attained by grinding. The problem was solved electrically. The coupling between the crystal and the tuned circuit behind the transmitter tube was made variable. The crystal shown in the circuit of Fig. 10 is electrically nothing but a tuned circuit, and this crystal circuit is coupled electrically to the circuit of the tube assembly LC by means of a tube. The self-excited natural oscillation may be changed to a small extent by varying the coupling between the crystal and the tuned circuit in such a system. This variation in coupling is here brought about by means of a small variable condenser C_1 in the circuit element AB. The wave length ($\lambda = 500$ m.) is

*Translator's note: From within 1 part in 100,000 to 1 part in 1,000,000.

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changed by 30 cycles by varying C_1 from 4 cm. to 40 cm. in capacity. The variable capacity can be replaced by the capa-

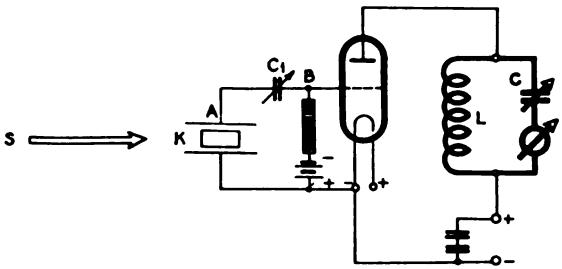

Figure 10—Variation of Wave Length by Means of C_1.

city of the electrodes; i.e., by varying the capacity of the electrodes by change of electrode separation, due to lifting one of the crystal electrodes, the frequency changes through 150 cycles.

Increase of the crystal temperature offers a further possibility for changing the crystal frequency. If the crystal temperature is raised 10° by heating, in a hot box, the frequency of 600,000 cycles decreases by 60 cycles. The variation is approximately as follows:

T=20 to 30°— 60 cycles
50 to 60°— 90 cycles
70 to 80°—200 cycles
90 to 100°— 50 cycles

Experience must show us whether such crystal-controlled equal-wave broadcasting can be attained such that both transmitters will maintain equality of wave length over an extended period of time without extremely careful supervision.

A peculiar phenomenon was encountered in the study of radio-frequency excited quartz crystals. If we bring a crystal, which is longer in the direction of the optical axis than is usually necessary for the production of oscillations—e.g., a crystal of the dimensions 5 mm. (along its optical axis), 1.5 mm. (along its electrical axis), 3 mm. (at right angles to the optical and electrical axes)—, to its sharpest resonance by means of radio-frequency excitation, we usually do not perceive any movement. If we move the crystal somewhat, it begins to rotate and it continues to rotate with accelerating rapidity. The experiment usually ends with the

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crystal being catapulted out from its two electrodes to distances as much as ½ meter. The crystal rotation is based on two new phenomena; namely, a purely acoustic effect and a crystalline deviation (or anomaly) in the direction of the optical axis.

The acoustic phenomena are very obvious on a longitudinal-vibrating crystal of the dimensions 28×13×5 mm. It is excited at its most intense natural wave length (*λ= 3,080 m.*). Very intense air fluctuations appear at both surfaces (Fig. 11) at the moment of inception of oscillations, (and they continue as long as the oscillations are

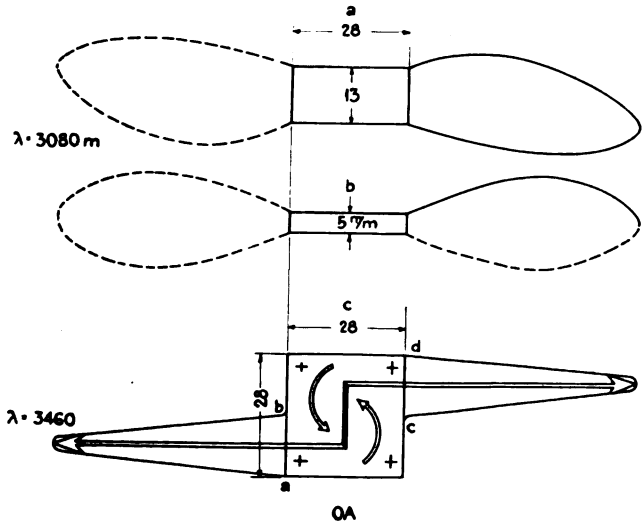

Figure 11—Air Fluctuation Near the Crystal.

sustained). If we drop lycoprodium powder in the vicinity of the surfaces it is immediately blown away. A candle flame is extinguished; a wind-actuated pinwheel, brought within the field of the air current, rotates. If we move the pinwheel towards the center of the air current, its rotation stops; moving the pinwheel in the other direction of the air current, makes it rotate in the opposite direction. Fig. 11a shows the distribution of these very intense air currents around the crystal in the horizontal direction, and Fig. 11b in the vertical direction. The limit of flow at the sides is quite definite.

We must delve into the reaction on the air of the high-

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frequency acoustic effects of the crystal for an explanation of these air currents. For these acoustic effects, the crystal oscillates as a longitudinal rod; i. e., the sound waves originate from its end surfaces. No air currents are produced by the sound waves alone, even for high frequencies, in the vicinity of the vibrating body. The sound waves merely produce equal condensations and rarefactions. The production of air currents must be explained by some anomalous relations. The very intense sounds set up, entirely disproportionate to the dimensions of the crystal, are anomalous. The crystal surfaces have the dimensions 13×5 mm. During condensation, the air particles are repelled either normal to, or at an angle to, the small surfaces. During rarefaction, the same air particles should move in the opposite direction towards the receding surfaces. Nevertheless, since the surfaces are small, and their perimeters relatively large, air particles can stream in from all sides during rarefaction. At the next outward impulse, there are, therefore, in addition to the former particles, all the particles which were added during rarefaction. The condensation and the rarefaction are no longer equally intense. The condensation overweighs the rarefaction. One experiences the sensation of a continuous current of air on account of the rapid succession of condensations.[1]

Of course, the crystal can not be set in rotation by these air currents and the forces of reaction accompanying them. The forces set up by the air currents vanish as far as their action on the crystal is concerned. A moment of rotation is needed to produce rotation. This we obtain through a crystalline anomaly in the direction of the optical axis. If we lengthen the crystal in the direction of the optical axis (Fig. 11c)—i. e., if it is 28 mm. instead of 13 mm. in the direction of the optical axis—,we find, on the one hand, that the air currents are no longer distributed over the whole of the side surfaces, but rather over half the surfaces; and, on the other hand, these air current surface areas are no longer symmetrical. The air current on the right side is towards the upper right of the side surface, that on the left side towards the lower left of that surface. No matter where the air currents are located, the sound produced is always the same for the same crystal.

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[1] Concerning ponderomotive effects of sound waves see Neesen, Annalen, 1887, p. 441; Davis, Sill. Journ. 1902, p. 129.

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A turning moment is exerted on the whole crystal by such dissymmetry of air flow.

The direction of rotation is the same as that of a Seegner waterwheel. If we reverse the crystal plate element, it rotates in the reverse direction because the air currents are attached to the surfaces *ab* and *cd*.

The rotation of the crystal can be used to form a small crystal motor. Fig. 12 shows an example. The crystal *(10 × 10 × 1.5)* is kept within bounds by a small spindle

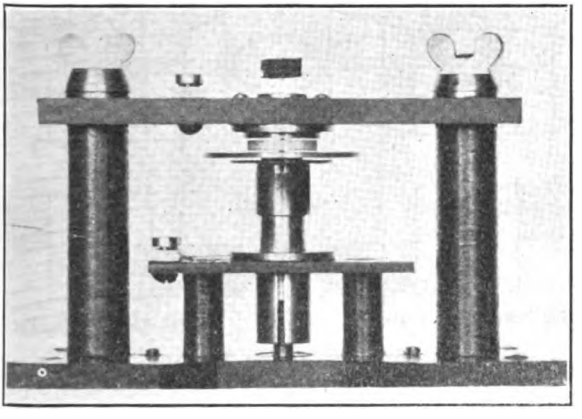

Figure 12—A Crystal Motor.

through its center. This spindle or shaft extends outwards and carries a small disk for ease in observing the rotation. As soon as the high-frequency source is connected, the crystal element begins to rotate; and stops when the source is disconnected.

A study was now made to see whether the electric-acoustic anomalies in the direction of the optical axis stood in any definite relation to the optical properties of the quartz. First, two similar elements of right and left optical-rotative quartz *(28 × 28 × 5 mm.*—Fig. 13b and Fig. 13d*)* were investigated. The electrical polarity of these elements, their electrical axes, were determined by first heating the elements to about 80° and dusting with red lead and sulphur powder during the process of cooling.* When the elements, placed between two electrodes with their negative sides uppermost, were excited at their most intense longitudinal os-

*The mixture should be shaken through muslin, when, by friction, the sulphur becomes negatively electrified and adheres to one side of the crystal, and the red lead, which is positively electrified, adhers to the other side.

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cillation, $\lambda = 3430$ *m.*, air currents arose, corresponding to those of Fig. 13b and Fig. 13d. Indeed, viewed in the direction of the electrical axis on the positive surface, the mo-

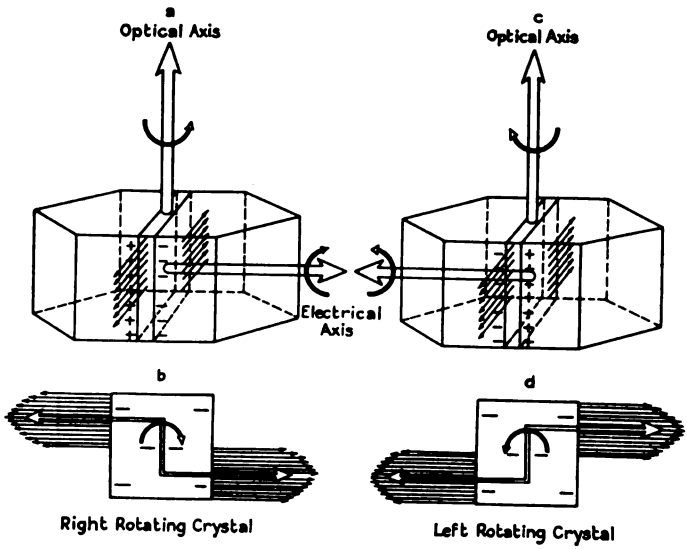

Figure 13—Right and Left Rotating Crystals.

ments of rotation of the air currents for the crystal rotating optically to the right are seen in the sense of a spatial rotation to the right; and in the sense of a spatial rotation to the left for the crystal rotating optically to the left. The optically to-the-right-rotative crystal rotates mechanically about the electrical axis to the right, while the optically left-rotative crystal rotates mechanically about the electrical axis to the left.

The direction in which the optical rotation of a quartz crystal takes place can, therefore, be determined from purely electrical and acoustical data without any optical study. A plate element parallel to the optical axis is cut from the crystal, the electrical polarity is determined by dusting the side with powder during cooling, and the element is excited at its longest and most intense natural wave length. If the air currents thereby produced are in the direction of a right turn about the electrical axis, the crystal rotates optically to the right. If the air currents of the crystal make a left turn about the electrical axis, the crystal rotates optically to the left. Fig. 14 shows figures of flow for the quartz plates of various widths, in the direction of the optical axis for

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quartz rotating to the right, and to the left, and for the fundamental and two harmonics.[1] Below, to the right, are given the diagrams of flow for a pair of crystals of the dimensions *5 mm.* $\times$ *7 mm.* $\times$ δ=*4 mm.* That side which is negative in cooling was uppermost for all the crystals. We see in all the figures that the crystals which turn to the right optically also rotate to the right when excited in their fundamental frequency, and vice versa. For the appreciably weaker harmonics of the crystal, the rotation is in the reverse direction to that of the fundamental, being again of

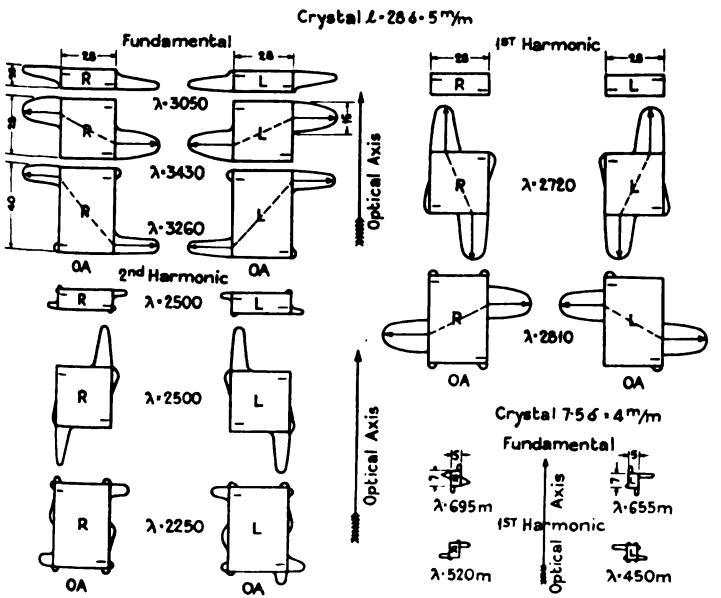

Figure 14—Right and Left Rotating Crystals.

opposite sign for crystals turning optically to the left and right. The line connecting the points of inception of the air currents at the crystal, "the angle of inclination," makes an angle of 40 deg. with, and to the left of, the optical axis for the main wave length of the right-rotative crystal. The "angle of inclination" of the first harmonic*, on the contrary slopes to the right, forming an angle of approximately 60 deg.—70 deg.

The lines, connecting the points of most intense sound production on both sides of the crystal for the fundamental

[1]The measurements were carried out by Dr. F. Michelson and P. Hagen.
*Translator's note. "Nearby vibration frequency" is a more correct interpretion of the German "Nebenwelle" than the term "harmonic".

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and for the first harmonic, point to the fact that there are surfaces across the crystal forming angles of 40° and 60° to the optical axis, such that the conditions for the propagation of sounds are especially favorable. If we assume as most probable that the surfaces of the crystal most favorable for the propagation of sounds are at the same time the surfaces of greatest molecular density in the crystal structure, these surfaces are indications of the manner in which the individual horizontal planes of the crystal structure are oriented. We thus have a new point of view for a conception of the crystal structure. The molecular density must be a maximum for those series of planes, the slopes for which we determined above. A structural model of the quartz element, which is based on these requirements, would have the following appearance. The structure-element consists of 3 horizontal planes. The planes are approximately equally spaced. In the planes are grid-molecule structures

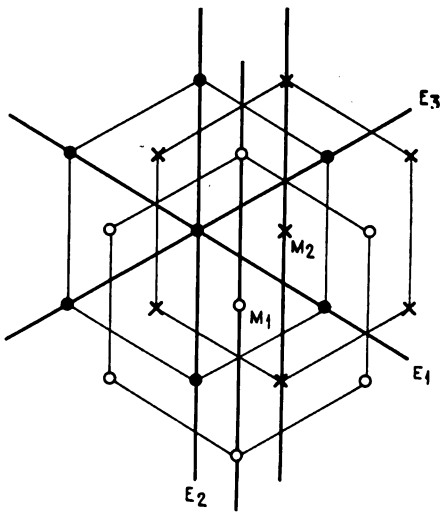

Figure 15—Displacement of the Crystal Planes.

consisting of equilateral triangles and hexagons, respectively. The dots in Fig. 15 correspond to Si-atoms of greatest atomic weight. The height (c) of the lattice structure is approximately equal to the length of a side of the triangle ($a=4.89$ A; $c=5.375$ A).* The 2nd and 3rd planes are displaced from the first plane, and this displacement is such that points M_1 and M_2 are displaced in the direction of the

*(Translator's Note: a equals 4.89 angstroms is evidently the length of a side of the equilateral triangle; while c equals 5.375 angstroms is the height of the lattice structure.)

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bisector of the angle between the electric axes E_1 and E_2, and E_1 and E_3, respectively, above the centers of gravity of the triangles that lie under them. The 4th plane is displaced with respect to the first plane by the length of a side of the triangle in the direction of one of the electrical axes.

It is thereby identical with the first plane. The left-rotative crystal is evolved from the right rotative one by exchanging the two planes 2 and 3. It is assumed in the model with respect to each Si-atom that there is an O-atom present above and below the Si-plane for each Si-atom. These atoms are intended merely to signify the direction of optical rotation of the crystal. The projection of the connecting line O-Si-O forms an angle of 120 deg. A rotation of 120 deg. is assumed from one plane to the neighboring plane—to the right for the right-rotating crystal; to the left for the left-rotating. This model corresponds to the following requirements:

Complete symmetry to all three electrical axes is maintained both for the right and for the left-rotating crystal.

There are planes of greatest molecular concentration. The angle of these surfaces to the optical axis is the same as was obtained from electric-acoustic observations—approximately 40 deg. These surfaces slope the same angle to the right away from the optical axis for the left-rotating crystal; and to the left for the right-rotating crystal.

There is another set of planes of largest molecular concentration (of the second order). The ratio of the molecular concentration of these planes to those of the first order are as 1:1.4. The surfaces are at a greater angle to the optical axis (60 deg.), and slope towards the other side of the optical axis from that to which the fundamental slopes. These surfaces correspond to the angle of inclination of the sound propagation for the first harmonic (λ=2,810 m.). The slopes of the planes, etc., of the left-rotating crystals are the reverse, as the corresponding sound diagrams require.

The translation of the mechanically right-rotating into the mechanically left-rotating system simultaneously changes the optically right-rotating crystal into an optically left-rotating crystal. Of itself, the optically right and left rotation of the polarization plane in the crystal has nothing to do with the mechanical rotation and the position of the surfaces of greatest molecular thickness. The optical rotation has to do with the atoms, the mechanical rotation with

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the structure of the space lattice of molecules. However, these quite independent phenomena are here rigidly associated. One rotation does not take place without the other.

If we cut a plate element out of the model, normal to one of the electrical axes, the element, in accordance with former conclusions, viewed in the direction of the electrical axis with negative plate uppermost, will be subjected to a moment of rotation to the right for the case of an optically right-rotating crystal with high-frequency excitation. We do not know the polarity of the plate; we know only the direction of surfaces of greatest molecular thickness, and know only that the corresponding crystal is, e.g., capable of rotation to the right. However, these two things suffice to define the polarity of the plate. For the slope of the surfaces of greatest molecular concentration to the optical axis is defined by these two items. The "angle of inclination" is translated into the optical axis by a rotation from the left to the right of a crystal rotating optically to the right. The negative face of the crystal ought to lie towards the spectator, according to previous statements with regard to simultaneously optical and mechanical right-rotating crystals.

The charges that go with the polarized condition of the faces are, therefore, fixed by the polarization rotation. One should progress from the position of the atom in one plane to the next in the structure element, for an elementary explanation of the appearance of the charge and its sign for quartz elements cut out at various angles. A point of reference is obtained from the atom positions along the axes of the structure element. By drawing the atoms of the three structural planes around one of the axes we get for each of the planes cut by the axis polarities differing from 30° to 30° on turning the plane.

In the construction of the above structural model the main stress was laid on the absence of contradictions to the results of electric-acoustic experiments. The model does not take into account the modern Roentgen-ray investigations. This method serves merely as a conclusion and may be regarded as a possibly useful aid to the understanding of the more specialized Roentgen-ray method.

The electric-acoustic excitation method may be applied even to non-piezo-electric crystals without much difficulty. Perhaps we may succeed in discovering some entirely new methods of gaining a conception of crystal structure.

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QUANTITATIVE MEASUREMENTS ON RECEPTION IN RADIO TELEGRAPHY[1]

By

G. Anders

(Abstract From Original Article Appearing in Elektrischen Nachrichtentechnik, Vol. 12, No. 2, 1925.)

In a recent Institute publication*, M. Baumler showed the results of a comprehensive series of measurements of American long wave stations, made by the National Telegraphic Engineering Bureau of Germany (Telegraphentechnische Reichsamt). Reference was made to the work of G. Anders who was responsible for the development of the system of measurement employed in obtaining these results. A complete description of this apparatus has been published in E. N. T.**, and the material is given here in somewhat abbreviated form.

Starting with a description of the simplest system of measurement in which the current in a loop is measured directly, the author gives a discussion of previous systems for the measurement of field intensities, their limitations and sources of error. Particular reference is made to the work of Hollingworth, Vallauri, Guierre, and Round in Europe, and of Pickard, Austin, Englund and Friis in this country.

The author has perfected a method of field intensity measurement in the German Bureau of Telegraphic Engineering. By this method a series of monthly measurements of several days' duration have been made of the field intensities at Berlin and Strelitz-Alt (Mecklenburg), and of the American Stations Marion and Rocky Point I and II†. This work started in February 1923, and was carried out with the cooperation of the Radio Corporation of America. The scheme employs the usual elements in signal strength mea-

[1]Received by the Editor, February 9, 1927. Presented before the Institute of Radio Engineers, New York, N. Y., March 6, 1927.

*M. Baumler "New Investigation Concerning the Propagation of Electromagnetic Waves." E. N. T. Vol. 1, Part 2, p. 50, 1924.

**Vol. 2, No. 12, P. 401, 1925.

†Proc. I. R. E. Vol. 13, No. 1, 1925 "Recent Investigations on the Propagation of Electromagnetic Waves" by M. Baumler.

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suring work, viz. wave collector, receiving set with indicator and calibrating apparatus. The basic arrangement is shown in Figure 6.

The calibrating method is essentially a current-measuring device. Referring to Fig. 6, the secondary receiving circuit is coupled through the mutual inductance M_1 to a small

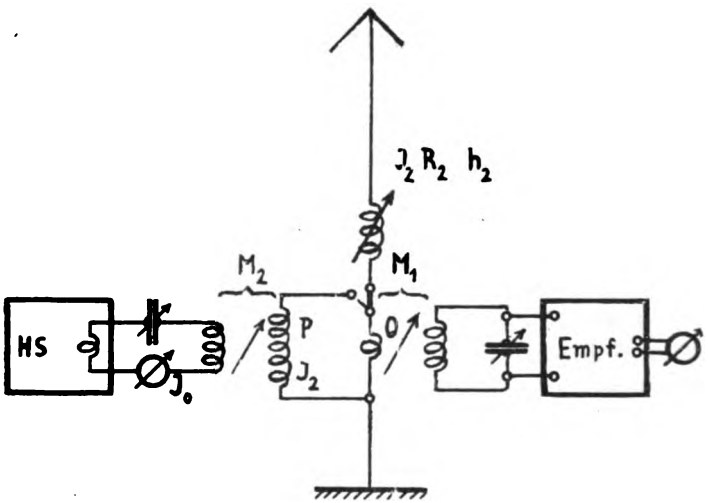

Figure 6

coil Q, which is alternately switched to the ground lead of the antenna circuit, and to the ends of a coil P of a relatively large number of turns, and of small effective resistance. This latter circuit is in turn coupled with the calibrating oscillator $H\ S$ by means of the mutual inductance M_2. These two coils may be considered as a current transformer. The primary current, I_o is measured by means of a vacuum barretter, (hot-wire resistance), and the secondary current, I_2, is made equal to the current produced by the receiving signal, as indicated by the meter in the output of the receiver. Then knowing the ratio of transformation of the current transformer it is possible to determine antenna currents which do not allow of direct measurement.

The use of a loop instead of the vertical antenna shown is equally permissible with this method, and it has been found possible to measure field intensities as low as 10 microvolts per meter at a wave length of 10 km. (30 kc.), using a 30-turn loop 3 meters square having an effective resistance of 10 ohms.

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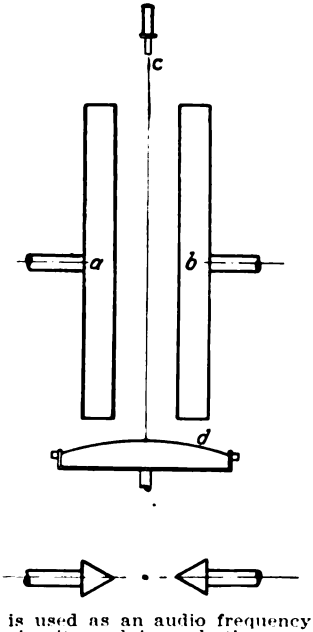
The receiving set proper consists of a two-stage radio frequency amplifier, heterodyne oscillator, detector, and an audio amplifier. A Wulf thread electrometer* with a transformer is connected in series with the telephone receivers to indicate the output. The thread of the electrometer is protected by the secondary winding of the transformer, because the short-circuit current of the transformer is too small to burn out the thread if it touches the electrode edges.

Fig. 7 shows the external appearance of a portable measuring set built in the workshops of the Telegraph Engineering Bureau, and capable of covering the range of 200-25,000 meters by means of interchangeable coils. Figure 8 and 9 give the scheme of connections with overhead and loop antenna, respectively.

*The single-thread electrometer produced by the Gunther and Tegetmeyer Company of Brunswick and herein illustrated has two sharp-edged electrodes **a** and **b** each 69 millimeters (0.236 inch) long between which is placed a platinum wire **c** 0.002 millimeter (0.00008 inch) thick. The edges are parallel and 12 millimeters (0.47 inch) apart. The platinum wire is stretched taut by a thin bowed quartz fiber **d** placed below the edges and parallel to the line joining them.



When the electrometer is used as an audio frequency indicator for receiver measurements, the circuit used is such that one edge is connected to the wire and the other to the case of the instrument. As a direct or alternating potential is applied to the thread (wire), it is attracted to one edge and repelled from the other. The deflection is observed with a microscope and measured by means of a micrometer eyepiece. The range of potentials covered by the electrometer with a normal tension on the thread is from 25 to 100 volts.

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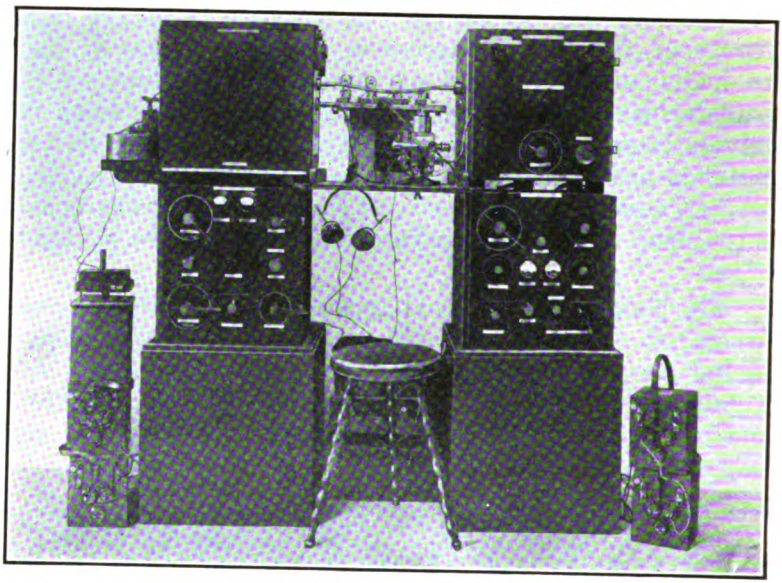

Figure 7—Portable Measuring Set

The apparatus consists of four square boxes, each 50 cm. on a side. The lower left box is No. 1 (see Figures 7 to 9). Box No. 2 is immediately above. The upper right box is No. 3. The lower right box is No. 4.

Toward the back of a shelf connecting the compartments

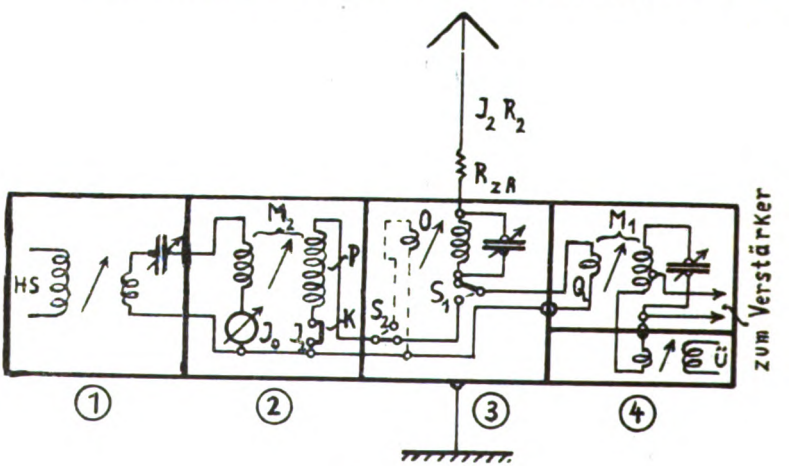

Figure 8

is located a two to four-tube audio amplifier with a resonant circuit below it, and an electrometer to the right.

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It was found necessary to protect the receiving circuit from the influence of interfering radio frequency fields. On the other hand, the difficulty of preventing reaction between oscillator and loop, especially at short waves, does not influence this method, since the aerial and oscillator act alternately on the winding Q of the receiver.

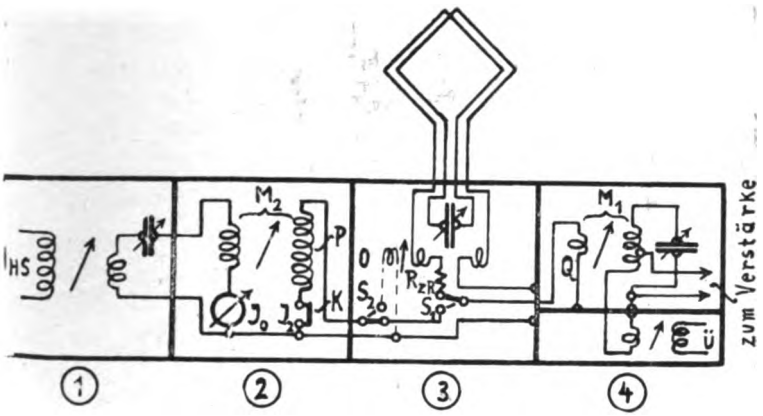

Figure 9

Stray coupling is prevented by the use of differential coils, and by enclosing each compartment in 0.3 mm (0.012) sheet copper, with appropriate interconnection of the circuit with this copper casing to minimize electro static disturbances. Interfering capacitative coupling within the compartments themselves is dimished by the use of symmetrical circuit connections.

The construction of the differential coils can be seen in Fig. 10. Each pair of coils consists of two pancake coils lying in the same plane and wound in the same sense, the current flowing in opposite directions. Differential variometers are used in the following parts of the apparatus:

1. The oscillator-measuring circuit coupling,
2. The current transformer,
3. Antenna loading coil and excitation winding (coil O), and
4. Coupling between coil Q and receiver coil.

The coupling between the oscillator and the measuring circuit is always made as close as possible, and the current in the measuring circuit is regulated by varying the oscilla-

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tor plate potential by means of a rotating contact form of potentiometer consisting of 25 steps of 400 ohms each. This keeps the oscillator power at a minimum and avoids excessive stray fields.

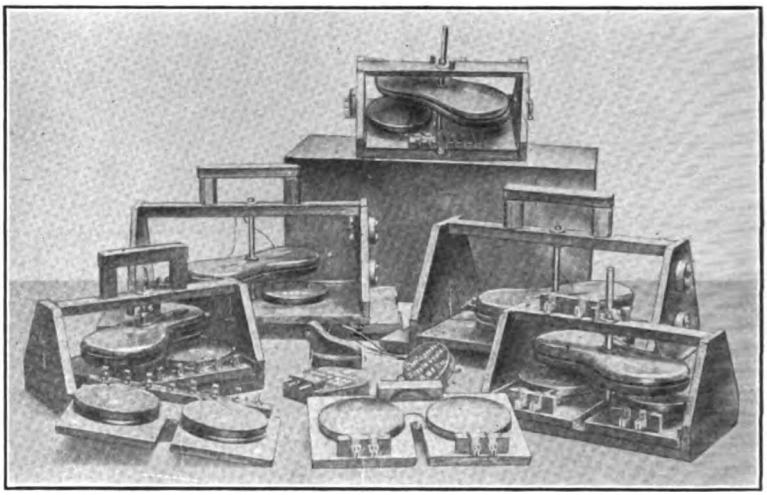

Figure 10—Construction of Differential Coils

In order to increase the sharpness of tuning of the receiver a low frequency anti-resonant circuit is connected in parallel with the primary of the transformer which couples the first and second stages of the audio-frequency amplifier. This tuned circuit consists of a 0.2-henry coil and a 0.1-microfarad condenser, and resonates at $\omega = 7{,}500$, which corresponds to the main resonant frequency of the telephone receiver used. The effective resistance of the tuned circuit was found by experiment to be 32 ohms, and the anti-resonant impedance 71,500 ohms. The sharpness of this circuit is often found to be too great when receiving short waves from stations with insufficient constancy of frequency, so that a step-by-step resistance is provided for adjusting the damping of this tuned circuit. In order to keep the impedance of this shunt constant, and so to keep the amplification the same, an appropriate value of resistance is inserted in series with the tuned circuit.

An important element in the calibrating apparatus is the current transformer. The theory of the current transformer is simple, being merely that of a transformer with-

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out an iron core. The current transformation ratio is given by the expression

$$\mu = \frac{\sqrt{R^2+(\omega L)^2}}{\omega M} = \frac{I_0}{I_2}$$

where R and L are the secondary constants, M the mutual inductance, and I_0 and I_2 the primary and secondary currents, respectively.

For a low-loss secondary coil R^2 is negligible as compared with (ωL), so that

$$\mu = \frac{L}{M}$$

It will be noted that this expression is independent of frequency. However, at the higher frequencies a dependance of μ upon frequency may arise as the result of capacity coupling between the coils, and the distributed capacity of the coils and leads. These effects may produce a decrease in μ. The relation is given by the expression*

$$\mu_1 = \frac{\mu}{1+\omega^2 K}$$

where μ_1 is the apparent ratio of transformation for a frequency factor ω, μ the ratio without capacity effects, and K a constant depending upon the magnitude of the capacity effect. K can be determined experimentally from two apparent ratios of transformation μ_1 and μ_2 at the two frequency factors ω_1 and ω_2,

$$K = \frac{\mu_1 - \mu_2}{\mu_2\omega_2^2 - \mu_1\omega_1^2}$$

(For a transformer in which $L=192\times10^{-4}$ henry, $K=1.76\times10^{-12}$ for $\mu=32.5$, and $K=2.73\times10^{-}$ for $\mu=2.03$. The useful range of this transformer was from $\lambda=6{,}000$ m (down.)

The tuned receiver circuit is coupled with the secondary circuit of the current transformer through the mutual inductance M_1 (Fig. 6). The receiver circuit induces into this circuit a resistance:

$$R_k = \frac{M_1^2\omega^2}{R_3}$$

in which R_3 is the receiver circuit resistance. But as long as $(R+R) << \omega^2L^2$, R_k does not affect μ and consequently the calibration.

*Moller and Schrader: "Concerning the Production of Small Alternating Currents of Known Amplitude", Jahrb. der drahtl. Telegr., 22, p. 64, 1923.

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A calibrated current meter is connected (in place of the short circuiting link, K, Figs. 8 and 9) in the secondary circuit for calibrating the current transformer. A sensitive thermocouple is used for the secondary and a less sensitive thermocouple or hot-wire meter for the primary according to the ratio of transformation to be determined. The current used will, of course, have to be larger than that used during measurements, so that larger tubes are employed. The ratio of transformation μ is independent of the current used because the transformer does not contain any resistance which is dependent on the current magnitude. The calibration may therefore be carried out at any arbitrary current values; in every case $\mu = I_0 \div I_2$, where I_0 and I_2 are the measured primary currents and secondary currents, respectively.

Ammeters which cannot be calibrated by connecting them in series with a d-c-calibrated radio frequency instrument can be calibrated very easily at radio frequencies by means of a calibrated transformer. The vacuum-barretter, the range of measurement of which -5×10^{-3} to 5×10^{-4} amperes—coincide with the range of sensitive vacuum-thermocouples only for its highest values, belongs to this class of current meters. The barretter resistance, which is highly variable with change in current (the resistance changes from 330 to 420 ohms as the current is changed from 0 to 0.4 milliampers), need not be considered in calibration so long as $\omega^2 L^2 > > R^2$. ωL must be at least 5000 ohms if R^2 is to be 1% of $\omega^2 L^2$, the barretter resistance 450 ohms, and the effective coil circuit resistance 50 ohms. (For $\lambda = 3{,}020$ meters, L must be at least 80×10^{-4} henrys).

Antenna calibration is required. It is necessary to determine the resistance of the overhead or loop antenna, in order to be able to ascertain the emf. induced by the received signal. This is accomplished by the substitution method with the adjustable resistances R_{za} and R_{zr} and the exciting coil O (see Figs. 8 and 9), using the measuring set as an indicator.

The disadvantage of the overhead antenna in measurement work is the necessity for determining its effective height. The effective height is dependent upon the wave length in question. It is constant only for wave lengths which are a multiple of the natural period of the antenna. The effective height may be determined:

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(a) By means of a sending station of known effective height h_1, and removed from the point of measurement by a distance of several wave lengths.

(b) With a loop set up at the same point as the overhead receiving antenna, or

(c) By the triangulation method of Pession.*

A discussion of these methods and the necessary formulas are given.

During the reception of signals, resistance is introduced into the aerial circuit as a result of its coupling with the tuned received circuit. The current flowing under the influence of an incoming wave is thus dependent upon the resistance introduced as well as upon the resistance of the aerial alone. Since in the calibration method a measured current is passed through coil Q, the introduced resistance must be known, or made negligible. It is satisfactory to make $M_1=1/10$ of the maximum value of M_1. There is then the advantage that it is unnecessary to take into consideration the induced resistance R_k, since it amounts to only one per cent. of R_2 for this case, while the emf. in the receiver circuit is decreased to but $1/5$ of the value obtained by using the maximum value of M_1.

The limits of the allowable values of M_1 are determined in practice by investigating the dependence of the electrometer readings on the angular position of the aerial-receiver coupling variometer, first with the aerial connection, and then with the oscillator connection. Before this is done the deflection produced by the oscillator with loose coupling is made equal to the deflection produced by a distant transmitter.

Receiver Requirements. Not every receiver is suited for this particular purpose. A receiver may be employed if it satisfies the following conditions:

1. The amplification must be sufficient and adjustable.

2. The deflection of the output-indicating device should be linearly related to the received signal intensity for any value of amplification.

3. The amplification factor should remain constant over long periods of time.

4. The selectivity should be as great as the constancy of frequency of the sending station permits.

*Radio Review, Vol. 2, p. 228, 1921.

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5. The apparatus should be as simple and as easy to inspect as possible, and as easily installed and tuned as practicable.

The receiving apparatus used consists of a two-stage radio frequency amplifier without regeneration, a heterodyne oscillator, an audio frequency amplifier with selective circuit, a thread electrometer, and a telephone receiver, all connected with a receiver circuit.

Three adjustments are necessary to tune the set-up to a given station:

1. The tuning of the antenna circuit,
2. The tuning of the receiver circuit, and
3. The tuning of the heterodyne oscillator, after first adjusting the antenna coupling to a suitable value.

All three tuning adjustments are effected by obtaining maximum electrometer deflection, and therefore can be more quickly and exactly carried out than adjustment for

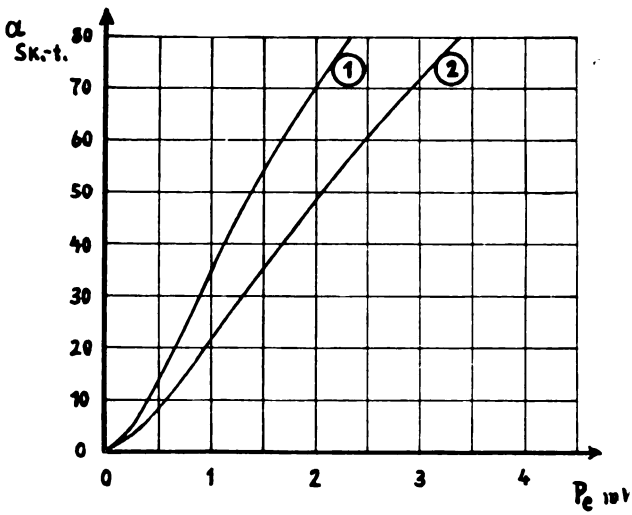

Figure 12—Electrometer deflection in scale divisions as a function of audio input voltage. (1) without resonant circuit, (2) with resonant circuit.

maximum sound intensity in the telephone receiver. The mutual independence of the three settings is of great importance in securing ease and rapidity of tuning. The aerial and receiver tuning adjustments are independent of one another because the coupling between the circuits is very loose. The heterodyne oscillator setting is not influenced

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by these adjustments because the position of the low frequency resonant circuit makes it independent of the input voltage of the audio amplifier. Without this tuned circuit this independence is absent because resonance effects, dependent upon the current through the audio transformers, influence the maximum electrometer deflections.

The proportionality of received emf. and electrometer deflection is of importance in the accuracy of measurement. The relation between electrometer deflection α and effective a. c. voltage P across the electrometer terminals, is almost exactly a parabola, and can be represented very closely, for normal electrometer thread tensions, by the equation:

$$\alpha = 5.9 \times 10^{-3} P^2$$

The relation between the current in the primary of the electrometer transformer and its secondary terminal voltage is approximately linear up to the greatest current values em-

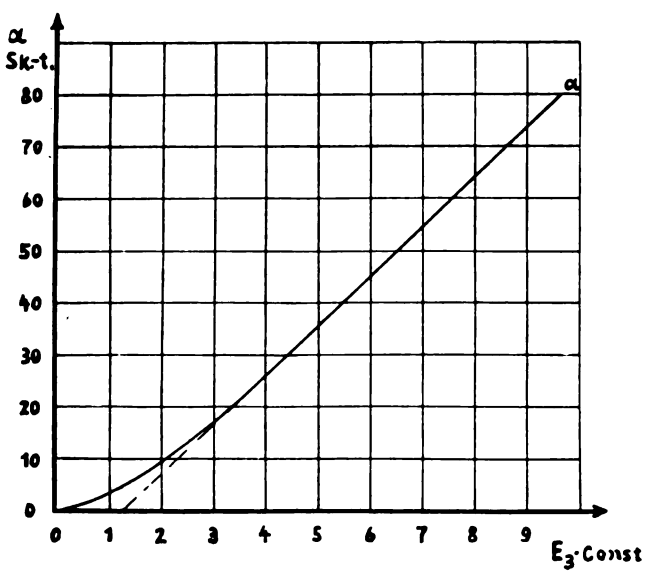

Figure 13—Electrometer response as a function of emf. across receiving circuit secondary.

ployed. The relation between electrometer deflection and voltage P_e applied at the input transformer of the three stage audio amplifier was arrived at experimentally: (1) without the resonant circuit, and (2) with the resonant circuit back of the first stage. Fig. 12 shows the two curves.

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The voltage amplification of the audio amplifier (ratio of electrometer voltage to the input voltage on the amplifier) is 77,500 without the resonant circuit, for $a=30$ scale divisions.

Fig. 13 shows the electrometer deflection a as a function of the secondary receiving circuit emf. E_s, for a wave length of 16,500 meters and for average set sensitivity. The curve is linear from 20 to 80 scale divisions. The relatively small initial sensitivity is unimportant since the deflection can always be brought up to at least 40 divisions by changing the combined amplification factor.

Tests of the Overall Accuracy of the Measuring Apparatus. The following tests were made to ascertain the accuracy of the results obtained with this apparatus. The transmitting current of one of the Königswusterhausen transmitters operating at a wave length of 7,200 meters, was varied from 4 to 29 amperes at intervals, and without the knowledge of the observer. Direct proportionality of measured signal strength to transmitting antenna current was observed.

A further proof was undertaken in June, 1923 in cooperation with the Nauen station using a 13,000 meter wave length. Since Nauen transmitted its normal power, sufficient signal intensity was received at Berlin to permit direct measurement with a simple collector and current measuring device. After making these observations the transmitter current was then reduced by a factor of 130, using a 10-watt oscillator to excite the aerial, and a measurement made with the present measuring equipment.

The per cent. deviation from the ratio of transmitting currents was + 4.29 % in the first case, and +4.76 % in the second. These errors represent the combined inaccuracies of all measuring devices at both transmitter and receiver, for the large and small intensities. This investigation served to show that the values of field intensity produced by American stations could be measured with this apparatus with an accuracy of at least ten per cent. Investigation has indicated that the average ratios of the received field strength and the product of ammeter indication by effective height of the Nauen station, as measured in the United States independently by Austin, Pickard, and Englund, were three to five times smaller than the corresponding ratios for the Marion and the Rocky Point I and II sta-

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tions, as measured in Germany, using the ammeter readings and the effective heights furnished by the Radio Corporation of America.

Comparison With Previous Methods

The use of a local calibrating oscillator is common to all methods for measuring weak fields. With methods in which comparison between the received signal intensity and the strength of the local oscillator is made by sound comparison with the telephone receiver, the optimum attainable accuracy is of the order of 10 % if the pitch and quality of audio notes remain the same. The accuracy drops markedly if the tone used as a basis of comparison is modified by non-uniform interfering sounds, especially if the signal is broken up by the rhythm of the Morse code, while the oscillator gives a uniform tone free from interference. This is the case for the methods employed by Round and by Austin. Ordinarily, the comparison oscillator intensity is set too low (up to 70 % error), as investigations of the author with the cooperation of non-partisan persons showed. The methods of Vallauri, Pickard, and Englund which employ a loop antenna also use a certain amount of interfering sound with the comparison oscillator, but it is possible that on account of the directional properties of the loop the interference emf. component perpendicular to the direction of reception does not affect reception.

Still more unsuitable is the "parallel resistance" method of comparison by means of a sound minimum, though this method is still used frequently for determining relative sound intensities with a telephone receiver. The highest attainable accuracy for complete absence of interference is $\pm$ 25 %. The errors which occur in practice are discussed by Dr. Baumler in the paper already referred to.

The thread electrometer used in this method as an indicator is so greatly superior to the telephone receiver that one might assume that it was unknown outside of Germany, or at least that it had not been employed as yet for the above purpose. On account of the small mass and the high natural period of the moving thread (wire), it is possible to tune all circuits quickly and exactly and to equalize accurately and reliably the energy of the oscillator without tiring the observer. The absence of interference in the calibrating

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source, a condition which leads to errors of measurement if the telephone receiver is used, actually allows of greater precision of adjustment when the thread electrometer is used. Moreover, outside interference is not marked by an increase in deflection, the suspension vibrating non-uniformly about the deflection which would be produced if interference were absent, so that one can distinguish between deflections caused by reception and by interference. The one disadvantage of the thread electrometer as compared with telephone receiver is the necessity for higher audio amplification, two more stages being required.

An added advantage of the Telegraph Engineering Bureau method is the possibility of making measurements with an overhead antenna without the use of an artificial antenna, the inexact determination of the constants of which might cause errors, and without the complications involved in Austin's method.

In the methods of Guierre, Vallauri, Pickard, and Englund (Friis) there occur difficulties in eliminating undesirable coupling of the oscillator and measuring circuits with the aerial or loop, especially with short waves. This defect is not present in the methods of the Telegraph Engineering Bureau and of Round, because the aerial or loop is disconnected when the oscillator is in the circuit.

Examined critically, it appears that errors may easily arise from failure to take into account capacitative coupling and the shunting effect of distributed capacity, if the mutual inductance of the aerial to measuring circuit is determined from the dimensions of the arrangement, as is done in the methods of Vallauri, Guierre, and Round. Still larger errors may arise from the use of resistance coupling between measurement circuit and loop.* (Methods of Pickard, Englund, and Austin.)

The current transformer principle used by the Telegraph Engineering Bureau is satisfactory, simple, and avoids the above mentioned disadvantages. Care must be exercised to see that the mutual inductance between the ground coil Q and the receiver circuit does not exceed the allowable values. Still, this need be set but once.

The difficulties encountered in the development of apparatus for measuring weak fields, particularly for short

*See the paper of Moller and Schrader previously mentioned.

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waves, are only partly evident from the above discussion, and only one who has carried out extensive researches in this field learns to know them all. The receiver part of the apparatus must satisfy much more rigid requirements than the ordinary receiver. The testing of a receiving set for suitability of result yielded cannot be carried out too carefully.

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THE FREQUENCY CHECKING STATION AT MARE ISLAND

BY

GEORGE T. ROYDEN*

Interference has always been one of the problems of radio communication, and it has become a serious problem, with the large number of high power radio stations now operating, to keep them from interfering with one another. To avoid congestion, a carefully planned allocation of frequencies to the several stations is made and measures taken to maintain the assigned frequency at each station. This paper describes the frequency checking established at Mare Island Navy Yard for the purpose of measuring the transmitted frequencies of naval radio stations extending from St. Paul, Alaska, to San Diego, California, as far west as Cavite, Philippine Islands, and as far south as Tutuila, Samoa. The frequencies transmitted during special test periods were measured at the Mare Island Radio Laboratory and the Officer-in-Charge of the transmitting station informed so that he might re-tune his transmitter in case of an error exceeding one half of one per cent.

The available methods may be divided into two general classes: individual control at the station, and supervisory control at a centrally located point in a limited region. The frequency meters at each radio station were not entirely satisfactory and subject to too great an inaccuracy to be suitable for the purpose of maintaining the frequency constant within the desired limits. Because of the limitations of equipment and difficulties of adequate coordination, the former method was considered less desirable than the supervisory control. The radio laboratory at Mare Island Navy Yard was designated to undertake this work in the Pacific Area. After considerable experimental work a method was devised which was sufficiently accurate, reliable, and sensitive enough to measure the frequency of the more important stations.

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A primary standard frequency meter of special design was constructed and apparatus purchased for calibrating it. This frequency meter consisted of two variable air condensers and a set of ten inductance coils. The condensers have rigid brass plates widely spaced and mounted within a metal case. The fixed plates are supported by three quartz pillar insulators. A vernier scale enables the scale to be accurately read to 1 part in 1800. The coils are wound with finely divided and insulated radio frequency cable on bakelite tubes 5 inches long and 5 inches in diameter. The table shows the number of turns and other pertinent data concerning these coils.

The calibration of the standard frequency meter depends primarily on the accuracy of a steel tuning fork, which had been carefully adjusted to 1000 cycles in com-

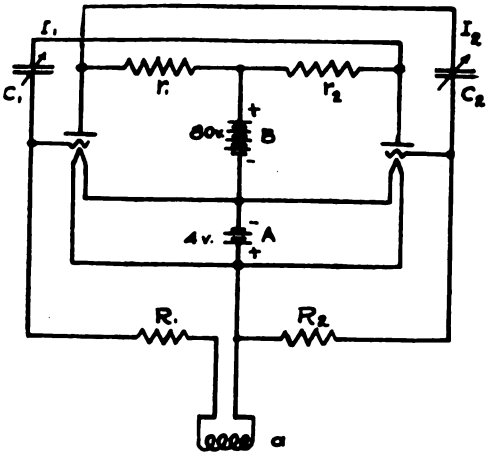

Figure 1—Multi-vibrator Circuit

parison with the standard clock at Paris. To avoid error due to a change in its frequency with temperature, this tuning fork was kept and used in a room, the temperature of which was the same as that when the fork was adjusted.

The comparison between the tuning fork and the frequency meter is made in a screened room by means of a multivibrator together with an amplifier and a heterodyne oscillator. The multivibrator consists of two vacuum tubes so connected, Figure 1, that a distorted alternating current is produced, illustrated in Figure 2. With such a peaked wave form, current will be sustained in a coupled circuit

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which is resonant to one of the harmonics. The set-up is made as indicated in Figure 3. After the apparatus has been operating long enough to maintain steady conditions, the multivibrator is set to 1,000 cycles, so that no beats

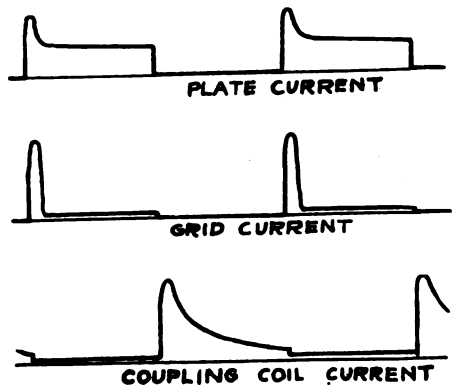

Figure 2—Multi-vibrator Wave Form

occur between it and the standard 1,000 cycle tuning fork. The frequency meter, which had been approximately calibrated in comparison with a separate standard, is set to the lowest harmonic within the range of the meter and the heterodyne oscillator adjusted, with close coupling, for a convenient beat note. The coupling is then reduced and the

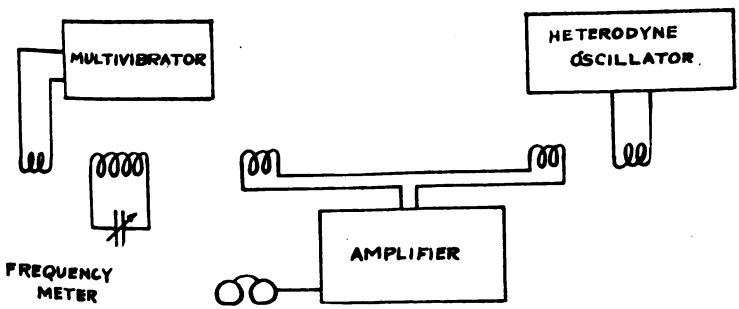

Figure 3

condenser varied for maximum response. This is repeated for the other harmonics, as many as 150 having been utilized without changing the fundamental frequency of the multivibrator.

This method gives a calibration point for each 1000 cycles, but with slight modifications, it is possible to obtain

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intermediate points. With close coupling between the heterodyne and multivibrator coils, the heterodyne is set at a frequency approximately midway between two harmonics, say the 14th and 15th. If set to 14,510 cycles there will result two audible beat frequencies, 510 and 490 cycles, which beat with each other. When adjusted so that this double beating does not occur, the heterodyne oscillator frequency is exactly 14,500 cycles. This can be transferred to the frequency meter by the grid reaction method or by noting the disturbance in the telephones when the frequency meter reaches resonance with the oscillator.

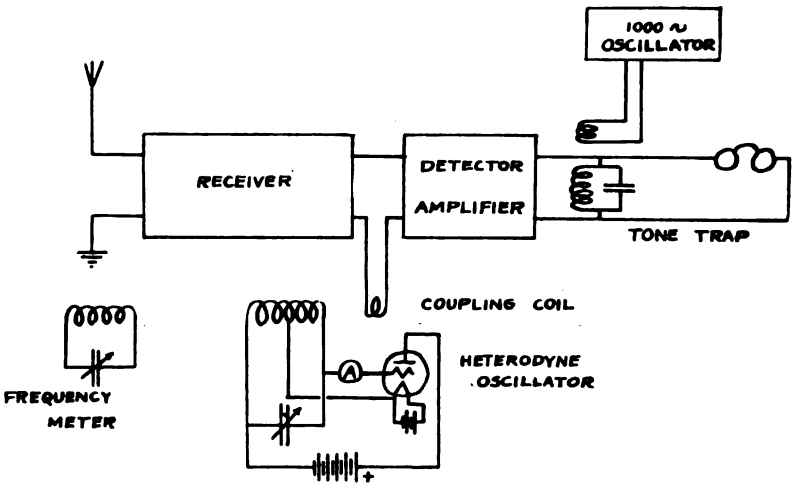

Figure 4

When this work was first undertaken, the frequency of a distant transmitter was measured by adjusting the heterodyne oscillator to zero beat note and then obtaining its frequency by the grid reaction method. This was not satisfactory because the silent period was fairly broad and easily obscured by interference, especially with weak signals. Another method used the sharply tuned standard frequency meter to increase the coupling between the primary and secondary circuits of the receiver. This was better but was abandoned in favor of a still more satisfactory method, in which the oscillator is adjusted to give a beat frequency of approximately 1,000 cycles which in turn beats against a 1,000 cycle tuning fork oscillator. The tone trap, comprising an inductance coil and condenser in parallel with the

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telephone receivers, serves as a means for adding the 1,000-cycle current from the tuning fork oscillator, increases the selectivity, reduces interference and facilitates adjusting the beat frequency to 1,000 cycles. Figure 4 illustrates the set-up. A very important advantage of this method is the check afforded by two independent observations, made with beat frequencies 1,000 cycles above and below the incoming frequency, which should be 2,000 cycles apart.

Another method is mentioned, originating in Europe and used at the Riverhead receiving station of the Radio Cor-

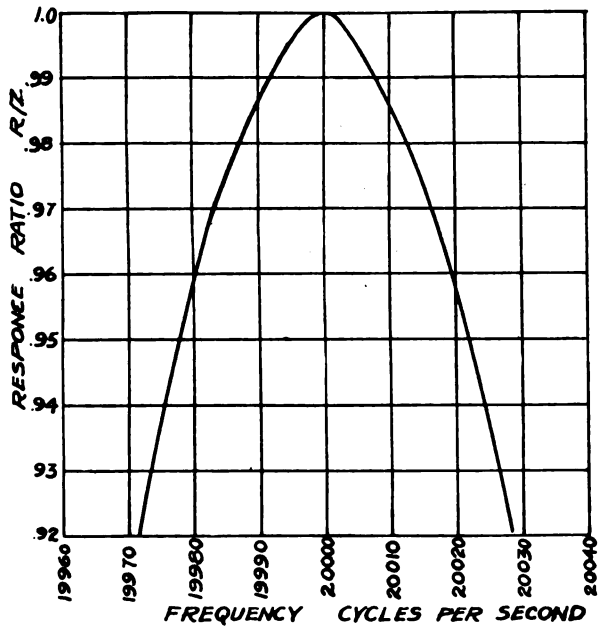

Figure 5—Frequency Meter Resonance Curve

poration of America, in which the frequency meter is connected with the coil and condenser in series as one arm of a bridge network, the other bridge arms being resistances. The signal is conducted to opposite terminals and the receiver connected to the other two. After tuning in the station the frequency meter is varied for minimum signal and the bridge arms further adjusted for a null point. A readjustment of the frequency meter balances the bridge network and indicates the desired frequency. Being a direct method, this permits great accuracy, but weak signals may be obscured by interference.

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The accuracy is dependent on the sharpness of resonance of the frequency meter and to some extent on maintaining the accuracy of the standard tuning fork. The resonance curve, Figure 5, with the coil for the 20-kilocycle range indicates that the band over which the frequency meter may be varied for a 1 % change is approximately 20 cycles. This should make it possible to determine the frequency within say 10 cycles for a single determination and say 8 cycles when several observations are averaged. Counting the calibration error it is not likely that the average of two observations will be in error greater than 15 cycles or better than 0.1 %. Repeated observations have demonstrated that such an accuracy with this method is easily obtained.

When measuring the frequency of a spark transmitter or a vacuum tube set with alternating current plate supply the 1,000-cycle oscillator is not used and the heterodyne beat note adjusted first to zero and then to beat with the modulation frequency, adjusting the oscillator frequency both above and below the incoming frequency, giving three observations, which are averaged.

This system has been in use for about five years, regular tests being held each month, and has been of material assistance in mitigating the interference problem in the Pacific region.

COIL DATA

PRIMARY STANDARD FREQUENCY METER

Coil	Wire Size No. B. & S	Turns	Inductance Hy.	Resistance D. C.	R.F.* Ohms
A	3x11x10 30	7	0.0000063	0.012	...
B	" "	11	0.000018	0.020	...
C	" "	23	0.000057	0.041	...
D	3x 3x16 38	39	0.00016	0.28	0.81
E	3x 3x16 38	65	0.00050	0.45	1.06
F	" "	110	0.0015	0.83	1.7
G	2x 3x16 "	198	0.0045	3.0	4.4
H	3x16 "	339	0.013	6.8	9.0
I	" "	559	0.041	12.6	15.5
J	" "	973	0.111	25.2	48.5

*Determined with a tuning capacity of 0.002 mfd.

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DISCUSSION ON THE OUTPUT CHARACTERISTICS OF AMPLIFIER TUBES (WARNER AND LOUGHREN)

E. Green: I have read with great interest the above paper in the Proceedings of the Institute of Radio Engineers for November 1926, as I have myself written an article on somewhat similar lines entitled "The Use of Plate Voltage Plate Current Characteristics in Studying the Action of Valves", which was published in "Experimental Wireless' for July and August 1926.

My object in writing to you is to point out that Captain H. J. Round has recognized the value of such characteristics for combining the characteristics of the valve and that of the output circuit to determine the behaviour of the combination. In the *Journal* of the Institution of Electrical Engineers for March, 1920, there is a paper by Captain H. J. Round on "Direction and Position Finding", and as Appendix 2 of this paper he dealt with the case of a simple resistance amplifier by means of plate current plate voltage characteristics. As far as I know this is the first published description of such use. In 1923 he explained the method to me, and realizing that it might be of value to others I set myself to write out, and to some extent apply, his method to various cases. The resulting article was accepted by the Editor of E. W. and W. E. in November 1924, but was not published until July and August, 1926. I am forwarding copies of this for reference.

DISCUSSION ON SIMPLIFIED S-L-F AND S-L-W DESIGN (O. C. ROOS)

R. R. Batcher:* It seems strange that an article containing this excellent analysis of S-L-F condenser design methods should contain such a statement as the following, "The S-L-F plate has its lower frequencies crowded together on the initial dial divisions." This statement is at variance with the very definition of an S-L-F condenser, the basic idea upon which this type of plate was initially advanced. This

*A. H. Grebe & Co. Inc.

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statement, however, is not new; it has been advanced for over a year as a sort of alibi by several manufacturers who have found that a true S-L-F condenser requires plates with a large swing so that it is difficult to manufacture accurately due to the large relative bulk for a given capacity. Before the advent of S-L-F condensers upon receivers, three years ago, the public had been used to having long wave stations separated at least five degrees for each myriacycle. After S-L-F condensers, all stations are separated equally, if true S-L-F operation is obtained. If this separation is sufficient for accurate manipulation of the dials on high frequency stations, it will always be enough on the low frequency stations.

It must be remembered that there is only a definite amount of dial space available. Assuming 100 dial divisions and 100 stations spaced equidistant on the frequency scale, there will be one station per degree. It is easy to set a dial manually to an accuracy of ¼ degree on any part of the dial.

Now assume that the hybrid condition of condenser plate as outlined (in Fig. 6 loc. cit.) has the same total area. Using the values set forth at the bottom of page 777 it is seen that we have crowded 80 stations in the S-L-F portion of the dial, i. e.—in approximately 42 dial divisions, or about one station every ½ division. We have actually thus crowded about 75 per cent. of the total number of the wave bands into half of the normal space. On the other hand we have allotted 105° to 28 wave bands or 28 stations to 58 dial divisions, or approximately one station to two divisions.

It can be stated that S-L-F design for condensers over the whole range is the only true and scientific method *as long as stations are maintained in the present bands.*

I devised a universal formula a number of years ago that will permit S-L-F operation over all or over any portion of the dial. It is necessary to know the same quantities outlined in the above article: the wavelength range desired (X) and the minimum capacity of the circuit (including that anticipated due to the condenser itself), also the maximum desired radius of the plates (ρ_m).

This formula is:—

$$\rho = \sqrt{\frac{K}{\left[1-\frac{\theta}{100}\left(1-\frac{1}{X}\right)\right]^2}+\rho_0^2}$$

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The formula for factor K has been worked out but in general it is found by substituting for $\rho =$ the maximum radius desired and $\theta = 100$ (note θ bears values from 0 to 100 corresponding to dial divisions) and solving for K.

Should occasion arise where S-L-F operation be desired over any portion of the dial it is only necessary to use the actual wavelength range for that portion in determining X and to assume θ refers to this dial section only.

It should be noted that it is to the designer's advantage to accurately determine the minimum capacity C_0 before starting, since a much greater plate area for a given rotor radius is obtained when C_0 is considered. Apparently a number of so-called S-L-F condensers were designed with this factor neglected, judging from the resulting plate shapes obtained.

O. C. Roos: Mr. Batcher deserves my thanks for "lighting" on the apparent slip—"The S. L. F. plate has its lower frequencies crowded together in the initial dial divisions". Of course this statement was meant for the generally manufactured "parts" condenser used by both radio telegraph workers and B. C. L. I meant to specify that when used at waves longer than 547 meters there would be crowding at "lower (than broadcast) frequencies etc". The intent is rather obvious in view of my analysis of S.L.W. design.

Regarding Fig. 6 of my paper this was in no wise intended for a practical case but simply as an illustration of the mechanism of designing procedure. Manufacturers and engineers nowadays often illustrate dimensions of their "machinery" without giving literal transfers of the constants used. The method is the thing and longer than broadcast waves were purposely included in Fig 6 to include "parts" condensers, which cover a wide range of use.

Mr. Batcher's formula resembles that of H. C. Forbes and like it fails to indicate the bearing of the physical factors by simple inspection. The ability to do this is the chief advantage aside from the saving of time inherent in my procedure and formulas.

Manufacturers who use gang condensers with uni-control must not use auxiliary condensers to correct coil errors and consequent failure to start off with all frequencies synchronized. The only possible remedies are either accurate duplication of inductances in all respects or else a reduction

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of range by turning the defective-stage condenser to a new starting point, thus reducing the range on the other stages. The auxiliary capacitances must not be touched however; since so-called "trimming" condensers, as such, are taboo. They "trim" the public—that's all.

As I am showing in other papers, digging into this subject, we must reduce the "separation" problem between stations to one of the relation between the dial detuning interval—D.D.I.—which reduces our resonance current by a given fraction—say 90 per cent., and the "dial spreading"—D.S.—which I define as the dial movement between stations divided by their frequency-change or practically the first derivative of dial movement in terms of frequency.

I have derived some interesting laws covering the four classes of commercial plates now used, S.L.F., E.L.W. (Exponential line wavelength, as in the Kolster Decremeter); S.L.W. and S.L.C.

Thus both the D.D.I and the D.S. vary as follows:

The D.D.I. and D.S. in S.L.F. are constant

The D.D.I. and D.S. in E.L.W. $\propto \lambda$

The D.D.I. and D.S. in S.L.W. $\propto \lambda^2$

The D.D.I. and D.S. in S.L.C. $\propto \lambda^3$

Here it is easily seen that the S.L.F. shape for broadcast frequencies which are equally spaced, is better than the famous E.L.W. or erroneously so-called Duddell "logarithmic" plate. Not only is the D.D.I constant all over the S.L.F. scale,—if the resistance is assumed by balancing of condenser and coil effects to be constant—but even if it varies as the frequency—as Bur. of Standards Bulletins indicate for some single larger coils,—the D.D.I. tends to a constant *percentage* of the scale reading throughout the scale and give what the E.L.W. could not give except by simultaneous coil and condenser variation!

Even if we could arrange our broadcast stations so that the next higher station frequencies were say, 102 per cent. of the last, we could not, employing a E.L.W. condenser and variable L, equal the performance of the S.L.F. plate with its practically fixed L. and fixed D.D.I. Many have dreamt about such a change but it has no practical foundation.

Note that all four types of condensers have with "flat" resistance characteristics, the same relative immunity from interference between stations. As the "Dial Spreading"

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varies so does the "Dial Detuning Interval" for all. Where they differ is in their power of spreading the stations themselves, so the manufacturer who wants to cover both the longer wavelengths and the broadcast band, must not only shift his coils, he must change his calibration. He will of course narrow his range and have only say 218 to 526 meters or even less.

It is easy to design an E.L.W. plate from a series of small S.L.F. or S.L.W. segments which step-by-step change their constants as we proceed from those covering the broadcast band to the longer waves. I have a paper on this subject which will shortly appear, but we "must break eggs to have an omelette" and our range must suffer as we "spread" stations. There seems no way out of it.

In conclusion, I believe Mr. Batcher's final paragraph is too meticulous. It should be precised as a well known first consideration among designers. He says "it is to the designer's advantage to accurately determine the minimum capacitance C_o before starting; since a much greater plate area for a given rotor radius is obtained—" This may with justice imply that the number of active dielectric "spaces" between rotor and stator should not be too great. We might leave too narrow a margin of constancy in C_o for the adjustable auxiliary calibration condenser (A.A.C.C.) to automatically compensate by its tested adjustment, the stray capacitances which may vary, through tube changes. Small plates tend in addition to a less constant value of stray flux with rotor movement than do large plates, but we have nevertheless assumed a constant stray flux, in all design methods.

In implying large plates he is adumbrating a critical condition of design when they become small; since for a given value of C_o and of calibration range, his number of dielectric spaces or double the number of rotor plates varies inversely as the square of his maximum radius i.e. reducing this from say 2.5 inches to 2.00 inches would increase the number of his rotor plates about 50 per cent. Machine shop studies have shown the best compromise between these factors, with results that dictate the proper range of values for C_o.

However, I may have misunderstood the more direct import of Mr. Batcher's last paragraph, which may mean that the "shape" of the S.L.F. or S.L.W. plate is dependent on

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something except the total turning angle (assumed 180 degrees or constant) and the range.

Assuming his "given rotor radius" to equal either the maximum or minimum radius permissible, his statement "that the designer should accurately determine the minimum capacitance C_0 before starting; *since* a much greater plate area for a given rotor radius is obtained when C_0 is considered.....a number of so-called S.L.F. condensers were designed with this factor neglected, judging from the resulting plate shapes obtained", is not a final criterion of design; since to assume a maximum rotor radius is the only safe procedure and will give a perfectly definite minimum or "zero" radius. If this radius is too small we must increase our maximum by redesigning. There's no other way out; as my range equations clearly show.

On the other hand if we follow Mr. Batcher's advice and start with the minimum or "zero" plate radius, in most cases we have too great a maximum radius and must redesign or use less than 180 degrees in our plate. The only difference lies in the fact that in the latter case we have fewer large plates and in the former more small plates, though the number of plates will not generally differ by more than 10 per cent. The question of the constancy of C_0 during the rotor movement determines the choice.

The latter part of his statement about the shape of the plates is erroneous if it implies that all actual S.L.F. segments for a given range are not "similar" geometrically, no matter what the value of C_0. Probably he referred to designs which showed no knowledge whatever of the effect of C_0 in the design, in which case I quite agree with him.

Paul M. Mueller: I have read Mr. Roos' paper on condenser designs appearing in the December issue and hasten to congratulate him on a very able presentation.

To augment his treatment of this problem I should like to call attention to some work on this same subject which I did about two years ago and which I published under the title "Tuning Tricks" in the August, 1926 edition of Q. S. T.

In this article, which was written primarily to show a method of properly balancing gaged multistage condensers, I brought out the fact that the polar area, approaching the origin and below zero of the dial must either be represented by stray capacitances in the circuit or by a small adjusta-

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ble auxiliary condenser. I then developed a design in which this neglected polar area was purposely made greater than the strays encountered in the usual circuit and outlined a graphical method for determining the proper adjustment of the auxiliary condenser to bring the total capacitance of the circuit up to the design value. Following this argument I was able to construct single control sets which functioned remarkably well throughout the entire change.

In this same article I indicated a method which is generally applicable for the design of condensers and which has some marked advantages, since determination of the plate generators follows mathematically when the five major considerations have been decided.

When one has decided upon

1. Minimum wavelength
2. Maximum wavelength
3. Type of tuning
4. Maximum stray capacity per stage
5. Thickness and type of dielectric

all of which are independent and empirical the following equations are applicable and lead to a concise result without recourse to "rule of thumb" methods.

Assume that the calibration must satisfy the condition that the rate of change of wavelength is proportional to the wavelength. (This calibration is intermediate between S.L.F. and S.L.W. and gives a more desirable station distribution than either of the more common methods.)

$\gamma=\gamma_o a^\theta$ satisfies the condition ..... (1)

Substituting and rewriting the terms of capacity

$C=C_o a^{2\theta}$ ..... (2)

The total circuit capacity C is made up of the geometrical capacity C_G plus the phantoms or strays and may be expressed thus:

$C=C_G+C_o$ ..... (3)

subst. in equation 2

$C_G=C_o a^{2\theta}-C_o$ ..... (4)

Now the geometric capacity is a function of the engaged area A of one plate

thus $f(A)=C_G$ ..... (5)

$$\left(\text{When } f=\frac{.2248\ (N-1)}{t}\right)$$

Subst. (5) in (4)

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$$A = \frac{C_0}{f} a^{2\theta} - \frac{C_0}{f} \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

$$\text{But } A = \frac{1}{z}\int_0^{\theta} (\rho^2 - r^2)\, d\theta \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7)$$

which is the general eq. for area in polar coordinates differentiating in terms of θ and combining

$$\frac{dA}{d\theta} = 2\frac{C_0}{f} a^{2\theta} \log_E a = \frac{1}{z}(\rho^2 - r^2)$$

or

$$\left(\frac{4C_0}{f}\right) a^{2\theta} \log_E a = (\rho^2 - r^2)$$

or

$$\rho = \sqrt{\left(\frac{4C_0}{f}\right) a^{2\theta} \log_E a + r^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (8)$$

If r or the inner radius of the rotor is constant as is usual we now have the outer generatrix in terms of the dial setting θ modified by the design constants phantom capacity C_0, dielectric f, and calibration a.

The shape of the condenser plate is now completely determined and we need only to design the inductance from the formula $\lambda_0 = f' \sqrt{LC_0}$ to fulfill the five major conditions and complete the problem.

It will be noted, since we have already considered the strays, that the numerical value of L is the pure or theoretical quantity which can be nicely established by mechanical measurement.

DISCUSSION ON THE CORRELATION OF RADIO RECEPTION WITH SOLAR ACTIVITY AND TERRESTRIAL MAGNETISM (G. W. PICKARD)

J. H. Dellinger: Mr. Pickard's paper marks a definite step in advance in our knowledge of the mechanism of radio transmission. It is not generally appreciated to what an extent this question of radio wave vagaries is the outstanding problem of radio engineering at the present time. It is not too much to say that this subject is the major and typical subject of the present era of radio development.

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Last week Dr. Pupin gave an address, as Retiring President of the American Institute for the Advancement of Science, on the subject of "Fifty Years' Progress in Electrical Communication." It is well for us to stop and realize that radio has been with us thirty of those fifty years. It was indeed just thirty years ago that the pioneer experiments of Marconi gave radio to the world. Broadly speaking, the radio era can be said to have had its beginnings at the commencement of this century, and the first three decades of the century are marked by distinct stages in its progress. The first decade (1900 to 1910) was the period of conquest of distance. Starting with a few miles and with the crude spark gaps and coherers of those days, the few pioneer radio engineers struck out in every direction for improvements, and by one ingenious means after another succeeded in steadily increasing the distance of communication. By the end of the decade radio had reached its goal of communicating to any desired distance on the earth's surface. There was a beginning of radio telephony but it was of a crude sort, and the real achievement of the decade was the establishing of radio telegraphy as a successful service.

The second decade was the period of development of the electron tube. Beginning with it as a low-power and imperfect device, it was steadily improved and made to handle higher and higher power. Its possibilities for the realization of radio telephony were steadily developed with the result that at the end of the decade everything was in readiness for broadcasting to make its appearance. This made radio an everyday necessity to a large fraction of the world's population.

Did it remain for the third decade of the century, the one in which we now are, merely to go on with minor improvements and refinements? Not at all. In spite of the perfection of transmitting and receiving apparatus, radio engineers and the public have been surprised to find that radio reception is far from perfect. We have no cure for fading, atmospheric disturbances, wave direction shifts, and other forms of interference and disturbances of reception. At the beginning of the decade we did not even know what caused them. There was very little information as to the laws of their behavior, much less of the laws of their production. For those scientists and engineers who are con-

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cerned with fundamental progress, this subject of wave vagaries was clearly a problem which had to be met and so it has remained.

Gratifying progress has been made and it may even be hoped that this decade may close with some generalization or climax of achievement in this field which will elevate radio to a new plane of service comparable with the arrival of broadcasting at the end of the last decade. The outlines of the mechanism of radio wave vagaries are now being discerned. It is still too early to give a positive explanation of the causes and characteristics of such things as fading and atmospheric disturbances, but there are a number of elements in the solution which are now well established. Numerous investigators have been doing pioneer work and assembling much valuable data giving the characteristics of fading, wave intensities, atmospheric disturbances, wave polarization, etc., as a function of various conditions such as time of day and year, frequency, weather, distance, topography, terrestrial magnetism, etc. Among the numerous things that stand out as a result of this work, one conclusion of interest, is that there is no important correlation between radio conditions and weather. Among the principal elements of the new knowledge are the role of the ionized upper portions of the atmosphere and the phenomena of very high frequencies. At the beginning of the decade the high-frequency phenomena were non-existent either in the activities or the thoughts of radio engineers. The theories of wave behavior were fairly simple and consistent for frequencies from the lowest to the highest then used. The work done during the past few years at frequencies from 2000 to 20,000 kilocycles has revealed a new world of wave behavior not fitting in with previous knowledge or theories at all. While introducing complications and inconsistencies with wave actions at other frequencies, the high frequencies have nevertheless introduced a vast amount of new information.

It is not my purpose in a discussion of this paper to give a summary of the knowledge of radio wave phenomena as it stands today. I would like to point out, however, that the correlation which Mr. Pickard has established is a major contribution to this picture of the radio wave mechanism which is now being assembled. How fruitful it will be to have this demonstration that the variations of electrical

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condition on the sun give rise to some of the characteristic radio wave variations, only time can tell. It can certainly be concluded that it will give a new impetus to further studies and analysis not only of radio wave phenomena but also the phenomenon of related sciences such as terrestrial magnetism and astrophysics.

DISCUSSION ON IMPORTANCE OF LABORATORY MEASUREMENTS IN THE DESIGN OF RADIO RECEIVERS

(W. A. MacDonald)

H. D. Oakley and **Norman Snyder**:* Mr. MacDonald says, "It is obvious that an exact knowledge of the individual and over-all characteristics of a radio receiver should be accurately known, yet experience shows that may manufacturers, including some of the largest, are practically unaware of the exact performance of the apparatus they produce".

We are ignorant of the facts upon which the second part of this statement is based, but agree, heartily, with the first part. Now inasmuch as Mr. MacDonald has put on record an accusation against many manufacturers of radio receivers and in his paper has dealt with the measurement of characteristics of individual units of a receiver, and not at all with the measurements of over-all characteristics of receivers, the manufacturers should be permitted, and should consider it their duty, to make a proper reply.

We should like to describe briefly the apparatus used by the General Electric Company for making tests on radio receivers, the quantities measured, and the method of test.

The testing equipment consists of two shielded rooms, a signal generator, voltage attenuating device, and an output voltmeter. In one of the rooms are the signal generator and input controls of the attenuator. In the other room are the attenuator, dummy antenna, (not used in the case of loop sets) receiver to be tested, and output voltmeter. These rooms shield both the testing equipment and the receiver from external disturbances and also prevent the signal generator from affecting the receiver except through the attenuator. The signal generator is a miniature broadcast transmitter employing the Heising system of modulation,

*Of the General Engineering Laboratory of the General Electric Company.

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and consists essentially of an audio frequency oscillator with a range of from 20 to 10,000 cycles, a radio frequency oscillator with a range of from 550 to 1500 k. c., a modulator, a radio frequency power amplifier, and a device for measuring the degree of modulation. The attenuator consists of a special form of inductor, the current through which can be adjusted to produce in the output of the inductor a known value of voltage. The output voltmeter is a device for measuring the effective value of audio frequency voltage existing across the output of the receiver.

Mr. MacDonald has described various tests made on individual parts of receivers, and since during the development of receiving apparatus this company makes similar tests, the tests described here will be those dealing with over-all characteristics, alone.

The tests and results to be described do not necessarily conform to our latest practice in investigating over-all

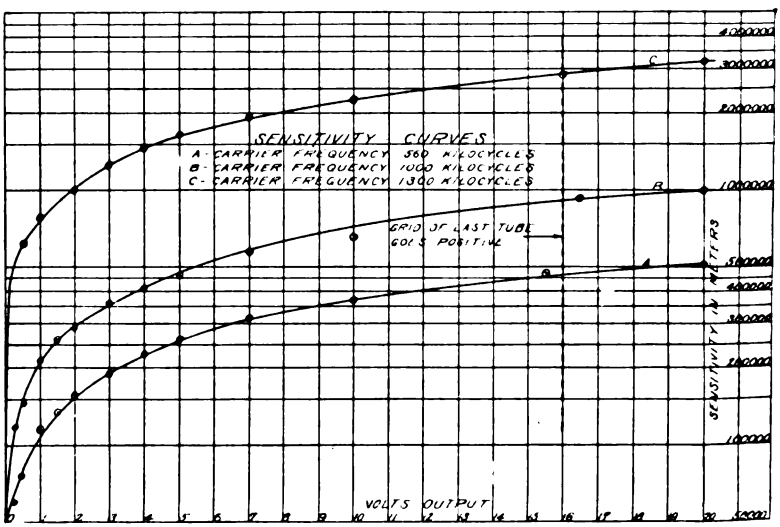
SENSITIVITY CURVES
A - CARRIER FREQUENCY 560 KILOCYCLES
B - CARRIER FREQUENCY 1000 KILOCYCLES
C - CARRIER FREQUENCY 1390 KILOCYCLES
GRID OF LAST TUBE GOES POSITIVE
VOLTS OUTPUT
SENSITIVITY IN MICROVOLTS

Figure 1

characteristics, but are representative of what has been and is being done in this class of measurements.

The quantities measured are sensitivity, selectivity, radiation, and quality. Figures 1, 2, 3 and 4 show the results obtained from tests made on a well known make of broadcast receiver. The circuit of this receiver consists of a

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two stage tuned radio frequency selector amplifier, detector, and two stages of transformer coupled audio frequency amplification.

If the receiver is designed to operate with an antenna, it is connected to a dummy having characteristics similar to those of an average broadcast receiver antenna, and in the ground side is inserted the output of the attenuator. If tests are to be made on a loop type receiver, the dummy, of course, is not used, but the low side of the loop is opened and the output of the attenuator inserted. The output of the receiver is connected to an RCA model 100 loudspeaker across which is the output voltmeter.

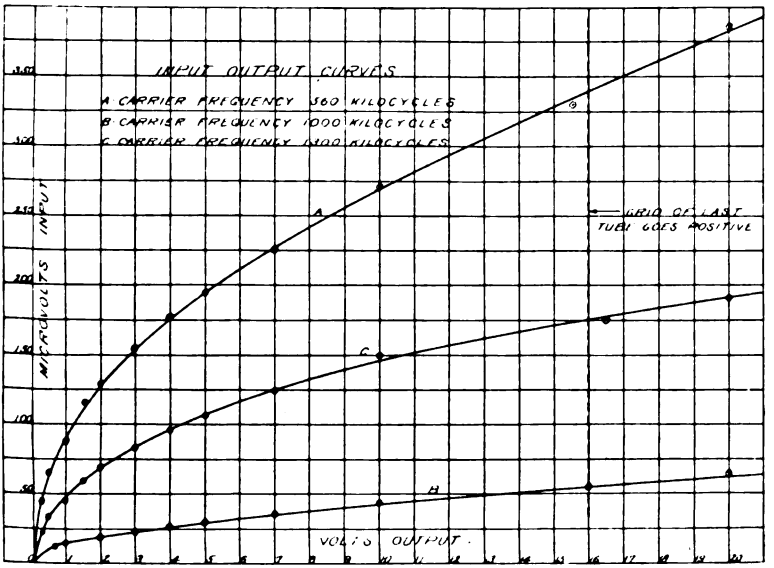
INPUT OUTPUT CURVES
A CARRIER FREQUENCY 560 KILOCYCLES
B CARRIER FREQUENCY 1000 KILOCYCLES
C CARRIER FREQUENCY 1400 KILOCYCLES
MICROVOLTS INPUT
GRID OF LAST TUBE GOES POSITIVE
VOLTS OUTPUT

Figure 2

Sensitivity measurements are made in this way. The receiver is tuned to some frequency within the broadcast band, and the signal generator is adjusted to give an output modulated 50% with a modulation frequency of 1000 cycles. The voltage induced in the input of the receiver (antenna of loop circuit) is increased in steps and at each step the corresponding output voltage is recorded. The test is continued until a point is reached where the grid of one of the receiver tubes (usually the second audio) starts to pass current. This is considered the operating limit of

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the set, since beyond this point distortion occurs. Dividing the voltage induced in the input by the effective height of the dummy antenna, or of the loop, the field strength is obtained. A plot of field strength against corresponding output voltage is one form of sensitivity curve. Another is sometimes preferred. This is a plot of sensitivity in meters against output voltage. The unit of sensitivity is called the meter and is defined as one volt output per one volt per meter input. Figures 1 and 2 illustrate sensitivity curves.

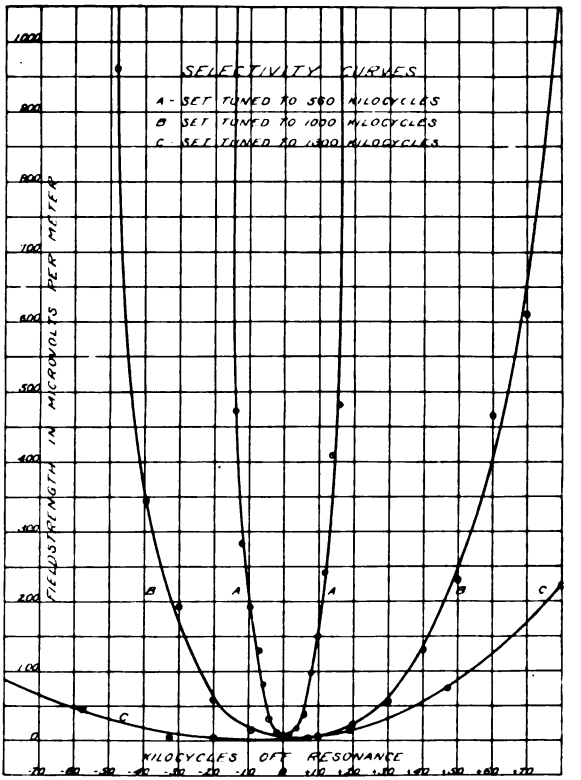
SELECTIVITY CURVES
A - SET TUNED TO 560 KILOCYCLES
B SET TUNED TO 1000 KILOCYCLES
C SET TUNED TO 1500 KILOCYCLES
FIELD STRENGTH IN MICROVOLTS PER METER
KILOCYCLES OFF RESONANCE

Figure 3

For selectivity measurements, the frequency of the signal generator is changed in small steps (maintaining the modulation frequency and degree constant) and at each step the frequency and also the input voltage, to receiver, required to maintain constant some value of output voltage are recorded. A plot of field strength against frequency is the selectivity curve. From it may be obtained directly the

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field strength an interfering station must have to produce a signal equal to the desired one. Knowing the response law of the receiver and the field strength of an interfering station, the signal strength may be indirectly determined. Figure 3 shows the selectivity curves.

The method of making radiation measurements is merely indicated here because a description of it would be rather long. The receiver produces an effect upon an indicator. The source of radiation is then replaced by a device having known characteristics, and a current flowing

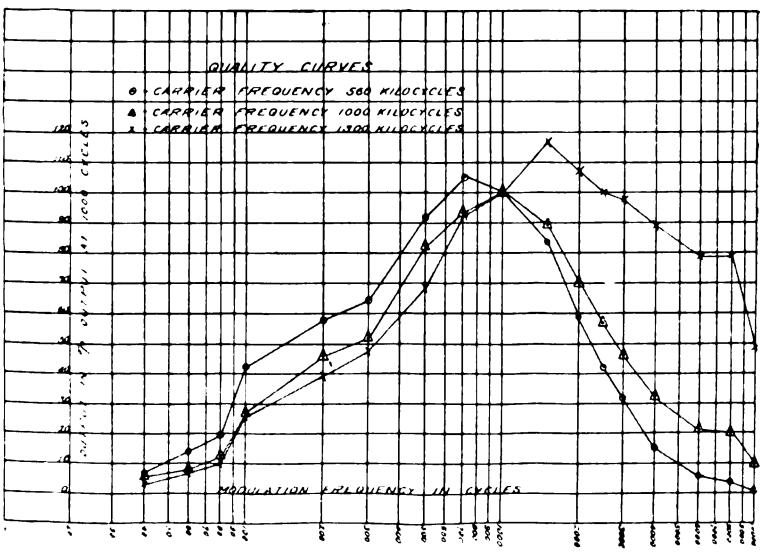

Figure 4

through it is adjusted so that the effect produced upon the indicator is the same as before. The radiation as determined from these tests is expressed in meter amperes. The particular receiver, whose characteristic curves are shown here, had no oscillator, nor could any of its circuits be made to oscillate, so of course no radiation could be measured.

The test for quality shows the discrimination of the receiver as a whole among the frequencies within the audible range, and does not show the presence of distortion due to the introduction of harmonics by the receiver itself. The test consists of maintaining a constant input to the receiver, at a constant carrier frequency and constant degree of

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modulation, but the modulation frequency is varied from 40 to 10,000 cycles and the output voltages at the different frequencies are recorded. The results are plotted as output voltage in per cent. of output voltage, at some one frequency, against frequency. The drooping of the low frequency part of the curve is practically all due to the audio transformer and loudspeaker characteristics. The drooping of the high frequency end is due partly to the detector circuit, partly to the radio frequency circuits tuning too sharply, and partly to some effect in the audio frequency amplifier. Figure 4 is an example of over-all quality characteristics.

J. H. Dellinger: The discussion of this paper at the Convention indicated some difference of opinion as to the relative desirability of overall measurements vs. measurements of components of the receiving set. Mr. MacDonald's paper omits overall measurements entirely and in justification of his position I wish to point out that his subject has to do with measurements for design purposes. There are at least three different purposes for receiving set tests, and it is altogether likely that the appropriate tests should be different depending upon which purpose is in view. These three different types of receiving set measurements would be those for—(a) the design of sets; (b) factory inspection tests of output; (c) tests suitable for the comparison of one receiving set with another. I assume that every receiving set manufacturer has an engineer whose duty it is to consider receiving sets from the last point of view. From the standpoint of the set user this is the most important type of receiving set test.

This subject is under consideration at the present time by the Receiving Sets Subcommittee of the Standardization Committee. This subcommittee has taken as its task of first importance the laying down of overall test methods, on the idea that methods which permit discrimination between the sets on the basis of final performance are most easily subject to standardization. While recognizing fully the difficulties of overall tests and the necessity of extremely careful specification of the conditions of operation of a set during such tests, it was nevertheless thought possible to render a useful service in standardizing testing methods by which one set could be compared with another. It seems

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much less likely that design tests or factory inspection tests could be standardized to advantage, since the designing engineer or production manager of any given manufacturer might be hampered rather than hindered by the endeavor to follow a standardized type of test. I wish to mention here the present ideas of the receiving set subcommittee in order to place the matter before the Institute for discussion and to gain for the subcommittee the advantage of criticism which any member may care to offer. I would be very glad to receive criticisms or suggestions from any Institute member or reader of the Proceedings.

The subcommittee has adopted tentative definitions of three overall characteristics, namely, sensitivity, selectivity, and fidelity. These are measured by the aid of a generator producing radio-frequency voltage of controllable modulation together with a means of measuring audio-frequency voltage across the output terminals of the set. In the following tentative definitions numerical values are to be inserted later for the quantities indicated by the letters *s*, *t*, *u*, *v*, *w*, *x*, *y*, and *z*.

Sensitivity.—Sensitivity is quantitatively defined as the ratio of the power output to the radio field intensity input measured as follows: Radio-frequency voltage, modulated *s*% at *t* cycles, is induced in the antenna if the receiving set includes one, otherwise in an artificial antenna consisting of a series arrangement of a capacity of *u* micromicrofarads, an inductance of *v* microhenries, and a resistance of *w* ohms. A curve is plotted of which the abscissas are the carrier frequency from 550 to 1500 kilocycles, and the ordinates are the radio field intensity in millivolts per meter required to produce *x* watts in a noninductive resistor connected across the receiving set output terminals (usually the loudspeaker terminals), the resistance of the resistor having been adjusted to give maximum output power. When the artificial antenna is used, radio field intensity is calculated from an assumed antenna height of *y* meters. The value of sensitivity for any one frequency is *x* divided by the ordinate of the curve for that frequency. The complete expression of sensitivity is the entire curve.

Selectivity.—Selectivity is expressed in terms of the frequency separation and the ratio of voltages produced across the output terminals, at and off resonance. It is not expressible by a single numerical value but requires one

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or more curves for its expression. It is measured by the same apparatus used for the sensitivity test. With the apparatus adjusted for the same input voltage, modulation frequency and percentage, and output resistance, as in the sensitivity test, the frequency of the source of voltage is varied without changing the receiving set tuning. The difference of frequency from resonance and the corresponding output voltage are observed. Pairs of curves are plotted, one of each pair for above resonance and the other for below resonance. Selectivity can be considered as expressed by any of three sets of curves, as follows:

1. Pairs of curves, taken at 550 and 1500 kilocycles and at any frequency between these two at which the selectivity is greater or less than at either extreme frequency, the curves having as abscissas the frequency separation from resonance in kilocycles, and as ordinates the ratio of output voltage across the output terminals at each frequency separation from resonance to the output voltage at resonance.

2. A pair of curves, one for above and one for below resonance, having as abscissas the carrier frequency from 550 to 1500 kilocycles, and as ordinates the frequency separation from resonance at which $1/z$ of the output voltage at resonance is produced by the same input voltage as produces x watt output at resonance.

3. That one of the two curves of paragraph 2 which shows the poorer selectivity.

Fidelity.—Fidelity is defined as the accuracy of reproduction at the receiving set output terminals of the modulation of the received wave. It is expressed by two or more curves, having as abscissas the modulation frequency from 30 to 10,000 cycles, and as ordinates the ratio of volts output at the modulation frequency of t cycles to output voltage at the modulation frequencey of measurement, both at $s\%$ modulation.

Harold A. Wheeler: Several of the points mentioned in the above paper, or in the discussion which followed its presentation, seem sufficiently interesting to warrant further discussion.

Reduction of Sidebands in a Radio Frequency Amplifier. In a tuned radio frequency amplifier, the selectivity is sometimes so great that there is a reduction in the amplitude of the outer sidebands carrying the modulation at the higher

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audio frequencies. This effect is proportional to the ratio A_s/A_c of the voltage amplification of the sidebands to that of the carrier, and is equivalent to only a reduction in the degree of modulation at the audio frequency in question.

Take the modulated wave of voltage e in the antennas of the transmitting and receiving stations,

$$e=e_o(1+m\cos\omega_m t)\cos\omega_c t$$

$$=e_o(\cos\omega_c t+\frac{m}{2}\cos\omega_s' t)+\frac{m}{2}\cos\omega_s'' t),$$

in which $\omega_s'=\omega_c-\omega_m$ and $\omega_s''=\omega_c+\omega_m$

ω_m=angular frequency of modulation, and

ω_c=angular frequency of carrier.

In general, when the amplifier is tuned to give maximum response at the carrier frequency, the ratio A_s/A_c is less than unity, and there is a "symmetrical" phase displacement of the sidebands. That is, the two sidebands have equal but opposite phase displacements θ,$-\theta$. Then the wave form delivered from the amplifier is

$$e'=e_o[A_c\cos\omega_c t+A_s\cdot\frac{m}{2}\cos(\omega_s' t+\theta)+A_s\cdot\frac{m}{2}\cos(\omega_s'' t-\theta)]$$

$$=A_c e_o[1+m'\cos(\omega_m t-\theta)]\cos\omega_c t$$

where $m'=m\cdot A_s/A_c$.

Therefore the principal effect is a reduction of the degree of modulation.

This is not accentuated in a detector whose response is proportional to the square of the applied voltage, since the modulated response is then proportional to the product of the carrier and sideband amplitudes, not to the square of the sideband amplitudes.

The phase displacement of the modulation is not detected by the ear, but is interesting in that it is the result of the "inertia" of an oscillating circuit. That is, there is an opposing reaction to any change in the amplitude of oscillation. For example, the time constant of equilibrium for oscillations in a simple series circuit (L, C, R) is $2L/R$. This corresponds to the time constant L/R for direct current in the simple inductive circuit L, R.

Reduction of High Audio Frequencies in a Detector. In the detector of Fig. 1, the grid condenser (C) requires some time to charge or discharge, which causes a discrimination against the higher audio frequencies of modulation. This effect can be reduced to a minimum by making C as small as possible without materially reducing the radio frequency

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voltage on the grid. Reducing the resistance $(1/g)$ of the grid leak has a similar advantage, but results in a material decrease in sensitivity to weak signals, accompanied by a material reduction of selectivity in the case of a tuned grid circuit.

It is generally immaterial whether the grid leak is connected to the positive filament terminal directly or through the coil, since any radio frequency current in the

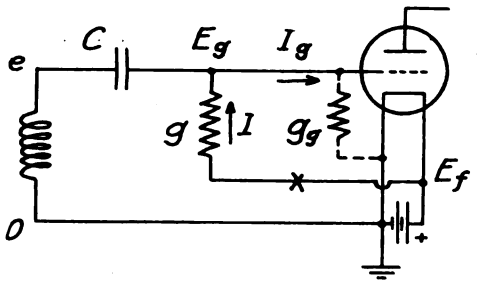

Figure 1

grid leak is ordinarily very small as compared with that in the tube grid conductance (g_g). There is no definite ratio between g_g and g, but one increases at about the same rate as the other.

The response of the detector to an unmodulated carrier can be explained by Fig. 2. In the absence of any signal, the

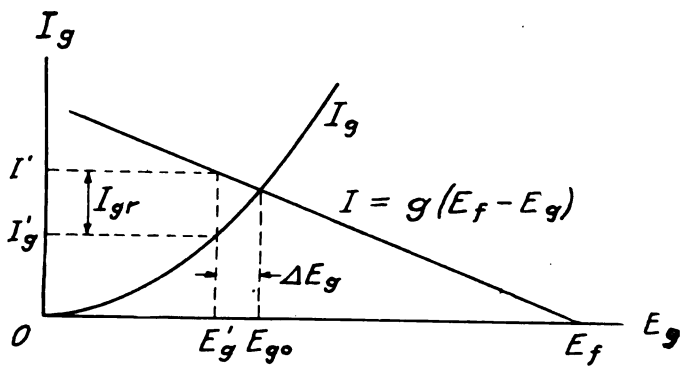

Figure 2

initial grid voltage (E_{go}) is determined by the intersection of the two curves, making $I_o = I_{go}$. When the signal is applied, the rectified component of the grid current (I_{gr}) requires that the grid voltage shift to E_g, such that

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or the equivalent $\Delta I=\Delta I_g+I_{gr}$.

$$I'=I'_g+I_{gr},$$

$$I_{gr}=\Delta I-\Delta I_g=-\Delta E_g(g_g+g).$$

Therefore $$\Delta E_g=\frac{-I_{gr}}{g_g+g}.$$

If e_g is the applied radio frequency voltage on the grid, then to a first approximation

$$I_{gr}=\frac{e_g^2}{2}\cdot\frac{d^2I_g}{dE_g^2}=\frac{e_g^2}{2}\cdot\frac{dg_g}{dE_g},\quad \Delta e_g=\frac{-\frac{e_g^2}{2}\cdot\frac{dg_g}{dE_g}}{g+g_g}.$$

This solution fails for a strong carrier, since g_g is defined only for small grid voltages. The behavior with a strong carrier is essentially the same, however. The conclusion is that for great sensitivity, (g_g+g) must be made small, which is secured by increasing the resistance of the grid leak.

When the carrier is modulated, the variations in the rectified grid voltage (ΔE_g) are accompanied by currents in C, as well as in g_g and g. The total grid admittance,

$$\left|g_g+g+j\omega_m C\right|=(g_g+g)\sqrt{1+\left(\frac{\omega_m C}{g_g+g}\right)^2},$$

takes the place of (g_g+g) in the denominator. The response to different audio frequencies of modulation is then proportional to

$$\frac{1}{\sqrt{1+\left(\frac{\omega_m C}{g_g+g}\right)^2}},$$

so that the discrimination against the higher frequencies is small if $\omega_m C<g_g+g$.

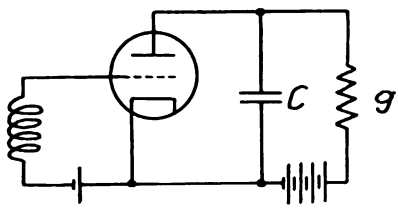

Figure 3

The magnitude of this effect is not easy to predict, because g_g is subject to so many conditions. Measurements of the relative response over the audio frequency range

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have been made, as mentioned in the paper, by introducing an audio frequency voltage at "X" in series with the grid leak. The resulting grid voltage then varies with the frequency in the same manner as it would with the frequency of modulation of a received signal.

The same effect is observed in the "plate detection" circuit of Fig. 3, with g_p taking the part of g_g in parallel with C. No general statements can be made as to the relative merits of the two detector circuits. The latter circuit is generally less sensitive, but the grid conductance is not present across a tuned grid circuit. In this case, g_p+g must be made small to secure great sensitivity.

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Review of Current Literature
Prepared by
STUART BALLANTINE
White Haven - - - - - - Pennsylvania

PROPAGATION OF SHORT WAVES AROUND THE EARTH

BY
E. QUÄCK

(Zeitscrift fuer Hochfrequenztechnik, 28, 177, 1926)

Some very interesting results have recently been obtained from observations of the commercial operation of short-wave transmitting stations, which furnish an important key to the propagation of radio waves of short-wavelength around the earth. In the month of October (1926) the Radio Corporation of America placed in operation the short-wave transmitting station 2XSA (now 2XT) with a wavelengths of 16.175 m. (18,550 kc), establishing communication with the German station Transradio A. G. für draht. Übersee-Verkehr (German Transatlantic Communication Co.). During the hours of communication the intervening space is in full daylight. It has been possible to employ signaling speeds up to and exceeding 80 words per minute. This communication, and also that previously established with Buenos Aires by means of the transmitter AGA (15 meters wavelength or approximately 20,000 kilocycles) at Nauen, has demonstrated the practical feasibility of daylight transatlantic communication with short waves. The American transmitter 2XT operates with about 12 kw. in the antenna.

In examining the autographic records of the signals received at Transradio Central, disturbing signals have often been discovered in spaces which one would normally expect to be blank. These may be interpreted as waves which have taken the other path around the earth. The signals of the American station have been photographed at the Transradio receiving station at Geltow by means of a Siemens oscillograph, and the records (see Fig. 1) clearly indicate the occurrence of a doubled-signal ("doppelzeichen") a short time later than the principal signal.

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These doubled-signals from 2XT are occasionally so strong as to interfere effectively with the operating records. Evidence of this is to be found in Fig. 2 which is a reproduction of a section of the recorder tape. During this test the American station transmitted the test letters *a,b,c,* and in the upper part of the record it will be observed that an extra dot appears to be added to the *b* (— . . .) thus converting it into a *6* (— . . . .); the lower record contains a distortion of the letter *c*. It is evident that we must shield our receivers from the subsidiary wave which encircles the earth in the other direction.

In order to make a more careful determination of the time which elapses between the incidence of the two signals, the American station was requested to transmit a special signal consisting of short dots at the rate of 5 per second. These signals were picked up by two short-wave receivers located at Geltow and each of the receivers was connected to a vibrator of the oscillograph. The first receiver was connected to a linear sloping antenna 20 m (65 ft) long and 15 m (49 ft.) high at one end; the second receiver was connected to a dipol (Hertz) antenna making an angle of 40° with the vertical and supported by a mast about 10 m. (32 ft.) high.

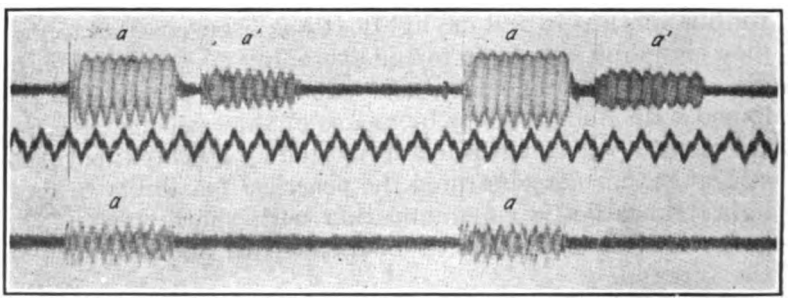

Figure 1—Doubled signals from the American short-wave station 2XT, λ= 16.175m. Photographed at Geltow, Oct. 11, 1926, 15^{m} h. (Timing wave 50 pps). a = principal signal; á = subsidiary signal.

The upper oscillographic trace in Fig. 1 shows the signal received by the first receiver, the lower trace that of the second; a 50-cycle timing wave appears in the middle. Unfortunately this timing current was not sufficiently constant. In more accurate measurements it will be necessary to use a steadier timing current of higher frequency in

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order to determine more precisely the small time-interval involved. Measurements of the upper curve show that the subsidiary signal lags behind the principal signal by about 0.096 second, hence it appears to traverse a path about 28,382 km. (17,600 miles) longer than that followed by the principal signal.

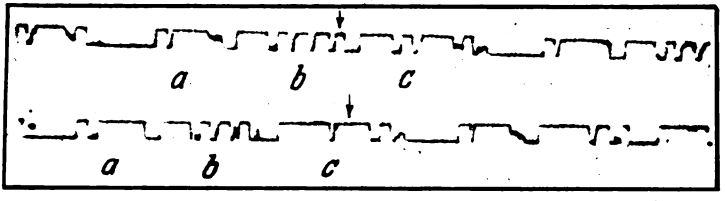

Figure 2—Automatically recorded signals distorted by doubling.

The oscillographic record of the second signal is also of interest in that it shows no trace of a subsidiary signal. This is probably because at the particular time of the photograph the dipol had an unfavorable position with respect to the polarization plane of the wave. It is likewise probable that the subsidiary wave had experienced fading; in earlier comparative observations of the kind, with less fading, no difference between the two antennas could be noticed.

A more accurate measurement with a steady timing wave of 1800 cycles was subsequently made (Fig. 3) which

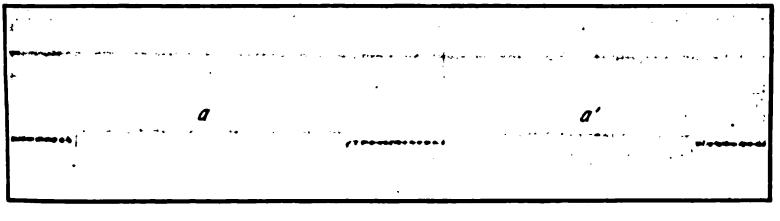

Figure 3—Doubled signals from 2XT, photographed at Geltow. Timing wave 1800 pps.

yielded 0.0857 sec. for the time interval, corresponding to a path difference of 28,705 km. (17,830 miles).

The foregoing results suggested a measurement of the signals originating at the station AGA (15 meters, 8 kw.) at Nauen. This investigation was equally successful. In Fig. 4 there is reproduced a section of the oscillograph record, in which will be noticed the incidence of the subsidiary

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signal after a time lag of 0.135 sec. Assuming a wave-velocity of 299,800 km./sec. the corresponding path-length turns out to be 41,499 km. (25,750 miles), which of course exceeds the circumference of the earth. (A still more accurate determination gave the result 41,200 km.). If it be assumed that this path corresponds to the circumference of a circle parallel to the great circle of the earth it appears from computation that the short wave has been propagated

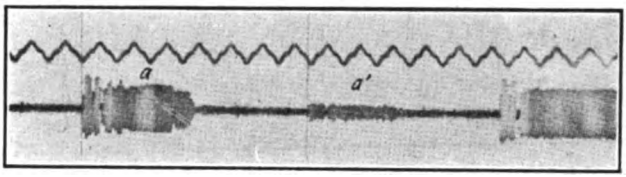

Figure 4—Photograph of subsidiary signal a' which had encircled the earth. Transmitter AGA (Nauen). $\lambda = 15$m., Oct. 20, 1926, 12^{23} h.

along a super-atmospheric stratum 182 km. (113 miles) above the surface of the earth. No proof that the wave actually takes this path is submitted, nor an explanation of how many such circular paths are utilized by the wave in returning to the receiver.

It may be remarked that these doubled-signals have been observed only in the case of very short wavelengths, 15—22 m. At longer wavelengths they are not registered, probably because their amplitudes have fallen too low. By suitable increase of the receiver sensitivity it should again be possible to detect a doubling on longer wavelengths. The method applied here should prove of service in further investigations of the processes of wave-transmission. It would be of particular interest to investigate the influence of the time of day and year upon the subsidiary wave. The preceding account is merely of operating experience and makes no pretense of exact measurement. It has at least unearthed the problem and shall leave its further investigation and explanations to more scientific methods.

SUMMARY

The autographic records of the signals received from the American short-wave station 2XT, $\lambda = 16.175$ *m.*, often contain disturbing signals the time lag of which behind the principal signal is such that they appear to be due to waves

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which have encircled the earth in the other direction. Such doubled-signals were oscillographically recorded and measured.

The signals from the station AGA (15 meters) at Nauen were also studied and a subsidiary wave which had traveled completely around the earth was likewise registered by the oscillograph. In this case the time interval was 0.138 sec., which with normal light-velocity gives a path of 41,200 km. This corresponds to a great circle girdling the earth at a height of 182 km.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED MARCH 2, 1927

Transferred to Member Grade

California	San Francisco, 219 Natoma St.	Heintz, R. M.
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New Jersey	Irvington, 219 Part Place	Rowe, C. M.
New York	New York, U. S. Coast Guard Barge Office	Fletcher, E. S.
Pennsylvania	St. David's	Stewart, Chas. H.

Elected to Member Grade

Georgia	Atlanta, Ga. School of Tech.	Shumard, Chas. C.
Louisiana	New Orleans, 109 Tchoupitoulas St.	Pendergrass, F. L.
New York	Rochester, 1060 University Ave.	McCanne, W. R.
	White Plains, 8 Church St.	Maloff, I. G.
Canada	Toronto, 90 Chestnut St.	Rogers, E. S.
Germany	Berlin-Sudende	Leithauser, G.
England	Bournemouth, Chintford	Everett, L. E. B.
	Stockport, Heaton Chapel	Hall, Albert.

Elected to Associate Grade

Arizona	Douglas, P. O. Box 1134	Schwamm, L. E.
California	Los Angeles, 3091 Pico St.	Burnell, J. H.
	Mountain View, State Highway	Winnegar, W. A.
	Pasadena, 53 N. Meredith St.	Nikirk, T. E.
	San Jose, P. O. Box 529	Cozzens, B.
Colorado	Pueblo, 1418 Bragdon St.	Bowen, C. L.
Connecticut	New Haven, 66 Norton St.	Conrad, V. A.
	New Haven, 66 Norton St.	Cordeman, W. P.
	New Haven, 66 Norton St.	Patten, S. F.
	West Hartford, 31 N. Main St.	Gade, W. H.
Florida	Key West, P. O. Box 201	Fernandez, M.
	Orlando, Western Union Teleg. Co.	Clark, M. D.
	St. Petersburg, P. O. Box 23	Rosseter, E. A.
	Winter Park, Box 344	Gree, C. M. Jr.
Illinois	Champaign	Hall, I.
	Chicago, 630 Roscoe St.	Adams, R. G.
	Chicago, 200 East Ontario St.	Brittin, F. L.
	Chicago, 3243 Congress St.	Callahan, J. J.
	Chicago, 4620 N. Leamington Ave.	Kablitz, A. P.
	Chicago, 1677 Milwaukee Ave.	Michael, C.
	Chicago, 851 Addison St.	O'Brien, W. J.
	Tuscola	Moulden, R. B.
	Urbana, 606 W. Elm St.	Morris, L. P.
Iowa	Dubuque, 1696 West 3rd St.	Short, D. W.
	Shenandoah, 205 W. Lowell Ave.	Rydberg, A. E.
	Villisca, Box 381	Harris, M. C.
Kentucky	Hopkinsville, Station WFIW	Harness, S. A.
Louisiana	New Orleans, 2222 Lapeyrouse St.	Galle, L. J.
	New Orleans, 4918 Saratoga St.	Gardberg, J.
	New Orleans, 1800 Broadway	Maier, O. J.
	New Orleans, 620 Fern St.	Moore, F. C.
	New Orleans, 511 E. Alexander St.	O'Neil, M. B.
	New Orleans, 1017 Peniston St.	Stevens, W. J.
	New Orleans, 5665 Rosemary	Zimbelman, F. A.
Maryland	Baltimore, 11 S. Gay St.	Ashbaugh, E. C.
Massachusetts	Boston, 186 Commonwealth Ave.	Cummings, L. G.
	Cambridge, 139 Mass. Ave.	Himoe, C. E.
	Newton Center, 25 Bothfeld Road	Miller, W. T.
	East Springfield, 20 Durgea St.	Hutcheson, J. A.
	Springfield c/o Westinghouse	Smith, H. Jr.
	Springfield, 20 Monrovia St.	Wells, D. W.
	Wollaston, 115 Clay St.	McGrath, J. L.
Michigan	Detroit, 2035 Calvert Ave.	McWood, W. L.
	Detroit, 1425 Park Place	Trowbridge, O. H.
Missouri	Kansas City, 4555 Main St.	Cook, Ellis C.
New Jersey	Atlantic City, 2707 Pacific Ave.	Flett, W. J.
	Bound Brook	Marks, E. R.
	Jersey City, 210 Jackson Ave.	Bremer, F. V.
	New Milford, Box 211	De Lime, T. L. Jr.
	Rutherford, 6 East Newell Ave.	Sargeant, C. E.
New York	Brooklyn, 214-6th Ave.	Gerel, A.

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	Brooklyn, 144A McDougal	Gibbons, J. J.
	Brooklyn, 132 Willoughby Ave.	Moxon, A. W.
	Hurley, Ulster County	Warren, W. A.
	Jamaica, 150-26 116th Ave.	Schwartz, N.
	New York, 233 Broadway	Bonett, D .J.
	New York, 233 Broadway	Cohen, C. M.
	New York, 227 East 239th St.	Howell, C. A.
	New York, 61 Cortlandt St.	Kendall, J. A.
	New York, 372 West 29th St.	Longboat, H. M.
	New York, 233 Broadway	Morehouse, T. B.
	New York, 233 Broadway	Smith, S. B.
	New York, 233 Broadway	Tunick, H.
	Rochester, c-o Stromberg-Carlson Mfg. Co.	Scoville, G. A.
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	East Plymouth, 45 West First St.	Budenkaye, A. C.
	Philadelphia, c-o Leeds & Northrup Co.	Doyle, E. D.
	Stonehurst, 7251 Guilford Road	Borkow, A. L. Jr.
	Wilkinsburg, 269 Halsey Ave.	Bock, A. P.
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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 15 MAY, 1927 Number 5

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GENERAL INFORMATION

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INSTITUTE ACTIVITIES

April Meeting of the Board of Direction

At the meeting of the Board of Direction of the Institute held at 4:00 p. m., on April 6, 1927, in the Institute Offices, the following were present: Dr. Ralph Bown, President; Frank Conrad, Vice-President; W. F. Hubley, Treasurer; Melville Eastham, J. V. L. Hogan, R. A. Heising, R. H. Manson, R. H. Marriott, L. E. Whittemore and J. M. Clayton, Assistant Secretary.

Upon recommendation of the Committee on Admissions the following admissions and transfers were approved: Transfer to the grade of Member: Pierson A. Anderson and R. S. Van Cleve. Election to the grade of Member: Stewart Aikenhead, Theodore C. Brown, Harold Buttner, Lieut. W. B. Goggins, W. H. Griffiths, Kenneth Harkness, O. T. McIlvaine, O. W. A. Oetting, Miles Pennybacker, L. E. Wells and Dr. Lewis M. Hull.

One hundred and forty-nine Associates and three Juniors were elected.

The Board voted that the Institute Medal of Honor for this year be awarded to Dr. L. W. Austin for his pioneer work in the quantitative measurement. The medal is to be presented at the June meeting of the Institute.

A petition from members residing within the vicinity of Cleveland, Ohio, for the formation of a Cleveland Section of the Institute was approved and the Section recognized.

New Institute Badge

To meet an insistent demand, the Institute badge is now available in the form of a watch charm emblem. It is of 14 karat gold, provided with a swivel ring for attaching to a watch chain or fob, and is finished on both sides. These watch charm emblems can be obtained from the office of the Secretary for Five Dollars each.

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New York Meeting

At the meeting of the Institute held on April 6, 1927, in the Engineering Societies Building, 37 West 39th St., a paper entitled, "Short Wave Commercial Long Distance Communication," by H. E. Hallborg, L. A. Briggs and C. W. Hansell, was presented by C. W. Hansell. In the discussion which followed Mr. Hallorg presented a portion of his paper, "Some Practical Aspects of Short Wave Operation at High Power." Both of these papers will appear in the June issue of the PROCEEDINGS. The discussion was participated in by many of those present.

The attendance at this meeting was over two hundred and fifty.

News of the Sections

Rochester Section

On March 4, 1927, a meeting of the Rochester Section was held in the Sangamo Hotel. Virgil M. Graham presided. The meeting was addressed by Dr. A. Hund, of the Bureau of Standards, on the subject of Piezo Electric Crystals.

This meeting was jointly held with the Rochester Section of the Optical Society. There were one hundred and twenty-five persons present.

The next meeting of the Rochester Section will be held on April 15, 1927, at which time a paper by Charles Bartlett, of the Weston Electrical Instrument Corp. will be read.

Canadian Section

The Canadian Section held its March meeting on the 2nd of the month in the Electrical Building of the University of Toronto. D. Hepburn presided. A paper by J. M. Thomson on "Audio Frequency Amplification with Transformer Coupling," was presented. A general discussion folowed.

The attendance was forty-six.

The April meeting of the Canadian Section was held in the Electrical Building, University of Toronto, on April 6th. D. Hepburn was the presiding officer.

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J. P. Henderson, of the Dominion Observatory, Ottawa, read a paper, "Radio in Surveying and Wireless in Time Signals." C. L. Richardson, A. M. Patience, V. G. Smith, F. K. Dalton and others discussed the paper.

There were eighteen members and fifteen visitors at the meeting.

Seattle Section

A meeting of the Seattle Section was held in the Telephone Building, Seattle, Washington, on March 5, 1927. John Greig delivered a paper on "Filtering Rectified A. C." The paper was discussed by Tyng Libby and others.

In the election of officers of the Seattle Section, which followed the presentation of the paper, the results were as follows: Chairman, Tyng Libby; Secretary-Treasurer, W. A. Kleist.

There were twenty-five members present.

The next meeting of the Seattle Section will be held in the Club Room of the Telephone Building, Seattle, on April 2, 1927. J. J. Ritter, of the Electro-Chemical Converter Company, will deliver an address on "Battery Eliminators for Radio Receivers."

Detroit Section

The Detroit Section held a meeting on March 18, 1927, in the Conference Room of the Detroit News Building. Thomas E. Clark presided. Three moving picture films, "The Audion," "The Electrical Transmission of Speech," and "The Story of a Broadcasting Station," were shown. A discussion participated in by Messrs. Hoffman, Buchanan, Glatzel and others followed.

In the election of officers which followed, Thomas E. Clark was elected Chairman, and W. R. Hoffman, Secretary-Treasurer. The regular meeting date of the Detroit Section was set for the third Friday of each month, except during July and August.

The next meeting of the Detroit Section will be held on April 15, 1927, in the Conference Room of the Detroit News. James McNary will deliver a paper entitled, "Measuring the Intensity of the Field Surrounding a Broadcasting Station."

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Los Angeles Section

On March 21, 1927, a meeting of the Los Angeles Section was held in the Los Angeles Commercial Club. L. Taufenbach presided. The address was by A. P. Hill, Supervising Radio Engineer of the Southern California Telephone Company, on "Recent Developments in Broadcast Transmission." The discussion which followed was general.

Cleveland Section

As previously announced, the formation of Cleveland Section has been approved. The officers of this Section are as follows: Chairman, John R. Martin; Vice-Chairman, R. E. Farnham; Secretary-Treasurer, L. L. Dodds.

Proposed Sections

Correspondence looking to the formation of Sections of the Institute is being held with interested Institute members in the following territories: Milwaukee, Wisc.; Schenectady, N. Y.; Pittsburgh, Penna.; New Orleans, La., and Minneapolis, Minn.

Committee Work

Committee on Admissions

At the meeting of the Committee on Admissions twenty-five applications for admission or transfer to various grades of membership in the Institute were considered. Twelve of these applications were approved, three were not acted upon for lack of sufficient information and ten were not approved.

Committee on Sections

The Committee on Sections, David H. Gage, Chairman, has been very active this year, having held four meetings since the first of January.

At present the Committee is working on the following matters: Preparation of a booklet of instructions to be used in the formation and management of Sections; Partitioning Sections into geographical districts by the Counties

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they include; A model form of Constitution and By-Laws for use in governing Sections; and establishment of new Sections in localities where there are a sufficient number of Institute members to maintain an active Section.

Committee on Membership

The Comittee on Membership, H. F. Dart, Chairman, has just completed a booklet describing the aims and activities of the Institute. This booklet will be useful in bringing the Institute before prospective members. It will be circulated to all Sections and other interested Institute members.

Committee on Publicity

The Committee on Publicity, W. G. H. Finch, Chairman, has regularly secured space in a number of newspapers announcing meetings of the Institute and reporting thereon. Much credit for the excellent attendance at the New York meetings of the Institute can be attributed to this Committee's work.

Papers in Pamphlet Form

The following papers are available in pamphlet form. Copies may be obtained free of charge by members by applying to the office of the Institute. The price to non-members is 50 cents per copy:

"Simultaneous Production of a Fundamental and a Harmonic in a Tube Generator," by Hoy J. Walls.

"Some Measurements of Short Wave Transmission," by R. A. Heising, J. C. Schelleng and G. C. Southworth.

"Reduction of Interference in Broadcast Reception," by Alfred N. Goldsmith.

"Dry Cell Batteries," by W. B. Schulte.

"A Theoretical and Experimental Investigation of Detection for Small Signals," by E. L. Chaffee and G. H. Browning.

"Vacuum Tube Nomenclature," by E. Leon Chaffee.

"Quantitative Measurements on Reception in Radio Telegraphy," by G. Anders.

"Piezo-Electric Crystal-Controlled Transmitters," by A. Crossley.

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"Piezo-Electric Crystals at Radio Frequencies," by A. Meissner.

Sectional Committee on Radio, A. E. S. C.

The activities of the Sectional Committee on Radio of the American Engineering Standards Committee for the past month have been as follows:

Component Parts and Wiring

A meeting of this technical committee was held at 29 West 39th Street, New York City, on March 2nd. The Chairman reported having received a list of N. E. M. A. standards for consideration by this Committee. This list of standards as well as vacuum tube standardization in general, was considered. It was agreed to defer standardization of vacuum tube bases until a report of the activities of the Vacuum Tube Committee has been received.

The standard A. E. S. C. vacuum tube socket markings were proposed and received unanimous approval of the committee members present. The Secretary has submitted this standard to the balance of the membership for vote.

The proposed date of the next meeting of this Comittee is April 6, 1927.

Vacuum Tubes

During the month drawings giving the exact dimensions and tolerances of the two forms of the "UX" base were received by the Chairman of the Committee, and were distributed to the membership for consideration for later standardization.

Electro-Acoustic Devices

No meetings of this Committee were held during February. A meeting is planned for the latter part of March.

Power Supply and Outside Plant and Transmitting and Receiving Sets and Installations

There were no activities of these Committees during the past month.

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Radio Manufacturers' Association Representatives

On vote of the Executive Committee of the Sectional Committee on Radio, an invitation to the Radio Manufacturers' Association to appoint a representative on all five of the Technical Committees of the Sectional Committee on Radio has been extended by the Secretary.

MEETING OF THE ELECTRO-ACOUSTIC DEVICES SUB-COMMITTEE

The Electro-Acoustic Devices Committee of the Institute's Standardization Committee held a meeting on April 5, 1927, in the offices of the Institute.

The following members of this sub-committee were present: R. H. Manson (Chairman), H. A. Frederick, F. C. Barton, C. R. Hanna, Irving Wolff, C. E. Brigham, Melville Eastham, A. Hund, B. E. Brown, Paul Andres and L. E. Whittemore. As guests the following attended: Messrs. Bostwick, McKown, Olney and Graham.

At this meeting forty terms and definitions were adopted.

These definitions include broad terms for the types of devices that are actuated by power from one system and supply power to another system, regardless of whether the systems are electrical, mechanical or acoustical. The word "transducer" has been taken as a fundamental term to designate a device of this type. By following this plan it was possible to build up a system of definitions that appear to cover the subject in a most complete form. For example, an electro-acoustic transducer is defined as a "transducer which is actuated by power from an electrical system and supplies power to an accoustical system, or vice versa."

The definition of a specific device such as a telephone receiver was built up in the following manner:

"A telephone receiver is an electro-acoustic transducer actuated by power from an electrical system and supplying power to an acoustic system, the wave form in the acoustic system corresponding to the wave form in the electrical system."

It was found necessary to refer to "wave form" in this definition in order that the scope would not be too broad and include such devices as an electric automobile horn.

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In order to differentiate between the various types of telephone receivers, reference is made to the acoustic load into which these devices operate. For example, a head receiver is defined as, "a telephone receiver designed to be fastened on the head of the user, and to operate into the ear as an acoustic load." On the other hand a loud speaker is, "a telephone receiver designed to radiate acoustic power into a room or open air."

In selecting terms to designate some of the particular elements of the electro-acoustic devices, it was found advisable to give serious consideration to present usage. Thus, the mechanical element that operates a telephone receiver or loud speaker is designated as a "driver element" and is defined as follows:

"The driver element of a telephone receiver is that portion of the receiver which receives power from the electrical system and converts it into mechanical power."

By using the former as a broad term, several terms to designate specific types were adopted, such as electromagnetic driver element, electrodynamic driver element and electrostatic driver element, etc.

It was considered that the scope of the Committees work included all electro-acoustic devices, therefore, definitions were adopted covering electro-mechanical devices, such as are now commonly employed in connection with electrical reproduction from phonograph records. After considerable discussion it was decided to use the term "phonograph pick-up" to designate the device that is actuated by a phonograph record and delivers power to an electrical system, the wave form in the electrical system corresponding to the wave form in the phonograph record. Thus, a "magnetic pick-up" is a phonograph pick-up whose electrical output is generated in a coil or conductor in a magnetic circuit or field.

One of the subjects that created considerable discussion in the committee is the proposed use of the "transmission unit" (abbreviated TU) in expressing the performance of electro-acoustic devices. This unit has been adopted as a telephone standard in practically all of the countries of the world and has already been used extensively in the radio industry. It was thought, however, by some members of the Committee, that the measurement of the output of a loud speaker was best expressed in dynes per square cen-

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timeter. For this reason further consideration is being given to the subject of Performance Indices.

NEMA HANDBOOK OF RADIO STANDARDS

The Second Edition of the Radio Division of the National Electrical Manufacturers' Association Handbook of Radio Standards is now available. This Handbook contains fifty pages of standards for many radio devices as adopted by the National Electrical Manufacturers' Association. Copies may be obtained at a dollar each from the Association at 420 Lexington Avenue, New York City.

Measurements at Radio Frequencies at A. I. E. E. Meeting

Information of very great interest to radio engineers will be presented in a symposium on measurements at high frequencies at the Regional Meeting of the American Institute of Electrical Engineers to be held at the Maplewood Hotel in Pittsfield, Mass., May 25-28. Members of the I. R. E. are invited to attend this symposium which will start at 10:30 A. M., Wednesday, May 25.

Fifteen papers are included in the symposium and they cover such subjects as radio-frequency voltmeters and current transformers; quantitative determination of radio-receiver performance; measurement of large high-frequency currents, radio field strengths, communication lines and apparatus, etc.

A complete program is published in the May issue of the *Journal* of the American Institute of Electrical Engineers.

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LOUD SPEAKER TESTING METHODS*

By
I. Wolff and A. Ringel

(Radio Corporation of America, Technical and Test Department)

A perfect loud speaker is one which when receiving perfect broadcasting through a flawless set will create the illusion of reality. Physically this means that the whole reproducing system should be able to transfer bodily a series of wavefronts without change from a certain region of the broadcast studio or concert hall to the listeners' room. The final test for the quality of a loud speaker is then the direct one of putting it on a receiving set, listening to it, and noting how nearly it fulfills this ideal.

For the person who wishes to develop and improve loud speakers this test is not sufficient. Neither broadcasting stations nor sets available are in the perfect class, and what sounds excellent on one combination may be very poor on another. The conditions given above can be more nearly satisfied by building a high quality speech amplifier for laboratory use. This is easily done with the aid of a condenser transmitter, resistance coupled amplifier, and a little equalization, the combination giving very nearly uniform response over the audible range. Precautions should also be taken to have the output tube of the amplifier capable of delivering the required voltage output without distortion. If someone talks into the transmitter in one room and a loud speaker is placed in an adjacent one, the listener in the second room can judge the quality of the reproduction. The weakest link in this chain of calibration lies in the sound pick-up. This problem is still far from solved in its acoustic aspects and of course until this is done the illusion of reality cannot be obtained. The value of this interesting test in the search for the perfect loud speaker, lies mainly in giving some uniform basis on which to compare models. For many purposes the best loud speaker is not the perfect one, but

*Received by the Editor March 21, 1927. Presented at a meeting of the Institute of Radio Engineers, May 4, 1927.

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the one which sounds best on the sets and broadcasting stations as they are.

In lieu of better apparatus, the speech amplifier may be used to determine what is wrong with a loud speaker which does not sound right. Resonance peaks stand out quite plainly and a person with a little experience can learn to detect and place them. Depressions in the response curve are not so apparent. It soon becomes obvious that some more quantitative method of making these tests is necessary

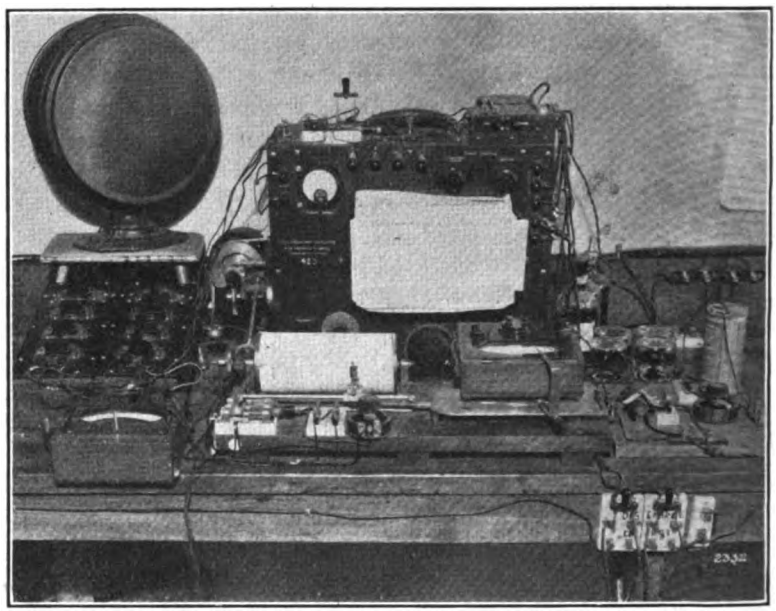

Figure 1—Oscillator and Recording Apparatus

if a real study and improvement of loud speakers is to be made. In order to find out the causes of certain defects in the response curve it is often necessary to know quite accurately the position and magnitude of the peaks and depressions.

The next obvious step is to use an electric oscillator of some kind giving as pure a wave as possible and impress its output on the loud speaker. The observer can get a pretty good idea, by listening, of the way the speaker responds to different tones in the audible frequency spectrum. It is however quite difficult to compare the intensities of tones of different frequency and the memory of the ear for tonal

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intensities which do not differ from each other greatly is poor. After an interval of a day or even an hour it is not possible to say whether a certain tone is more intense than the one heard previously. A test is required which will record absolutely the intensity of the tone.

We have found most satisfactory a system which makes a written record in two or three minutes of the loud speaker output measured in pressure of the sound wave. The apparatus used and the procedure followed will be described below under the headings—Oscillator, Sound Pickup and Recording Apparatus, and Position of the Pickup. A picture showing the oscillator and attached recording apparatus is shown in Figure 1.

OSCILLATOR

For making loud speaker measurements the most convenient electrical source will give a pure, simple harmonic output of constant voltage over the whole range to be

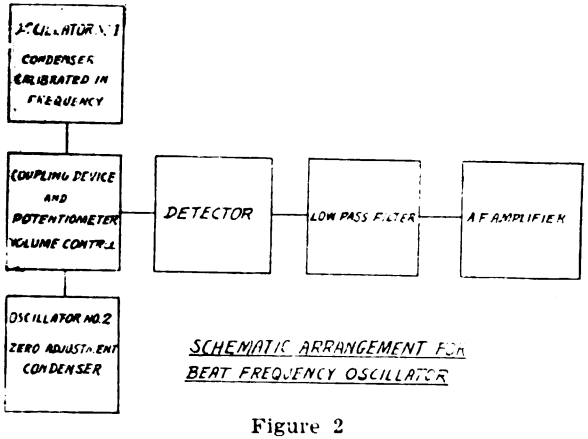

Figure 2

studied. It is also advantageous to have the frequency continuously variable rather than changed in jumps.

The above requirements are all reasonably well attained in an oscillator where the audible tone is the result of the beating of two tones above the audible range which differ somewhat in frequency. In the oscillator used in this laboratory one of the circuits is kept fixed at about 135,000 cycles while the other is varied by means of change in capacity to give the audible frequency wanted. One of these oscillating

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circuits has a fine adjustment for zero setting as will be described below. These two circuits are loosely coupled through a potentiometer volume control to a detector, which feeds through a low-pass filter with cut-off at 75,000 cycles to a two step resistance coupled amplifier. A low impedance tube is used in the last stage to give high power output. The oscillator is shown schematically in Figure 2.

A voltmeter placed in the output of the oscillator indicates that the output does not vary more than 5% from 30 to 10,000 cycles per second. A difficulty always encountered in instruments of this kind is the drift in frequency caused by changing filament current. This is reduced somewhat by having the two oscillating circuits as similar as possible. It is always necessary however to have some means for adjusting the zero point. If the frequency of both circuits remained constant for any particular condenser setting one of the circuits could be kept fixed and the other could have its condenser calibrated directly in frequency. Now suppose that the frequency of the calibrated circuit should change somewhat, due to a change in tube characteristics. The whole calibration would be thrown off by an amount almost constant over the whole range, and therefore if the other circuit had its frequency readjusted so as to make the calibration correct again at one point the whole scale would be correct. We therefore have the dial on the condenser in one of the oscillating circuits calibrated in frequency, while the other is used for zero adjustment. This adjustment can always be easily made by means of a tuning fork or by comparison with the 60 cycle power line. It should be made at some fairly low frequency so that if there is a non-constant shift in the calibration due to change in base frequency the per cent change in the audible frequency will be small.

The current to the loud speaker should be fed to it in such a way as to represent the output of a perfect set. Now in the perfect set the potential impressed on the grid of the last tube will be the same at all frequencies. Due to the fact that the ratio of the plate impedance of this tube to the loud speaker impedance is not always small or constant, the output potential applied to the loud speaker will not be constant. In order to represent the conditions of a perfect set as nearly as possible the constant potential should be applied to the loud speaker through a resistance equal to the

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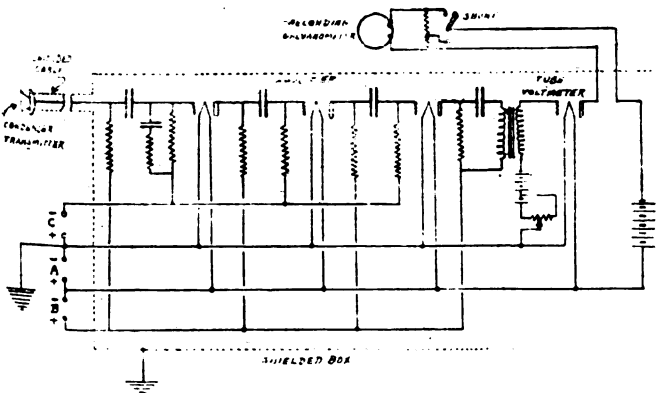
plate impedance of the tube with which it is to be used. This potential can then me measured with an a-c. voltmeter.

Sound Pickup and Recording Apparatus

The condenser transmitter developed by E. C. Wente[1] has been found most practical as sound pickup. One of these transmitters, calibrated by The Bell Telephone Laboratories, is used to feed into an amplifier made up of two resistance coupled stages and a third tube feeding through a transformer to a vacuum tube detector. The output of the detector is measured by means of a d-c. galvanometer. The condenser transmitter is ordinarily more sensitive at low and high frequencies[2] than in the middle range. The overall characteristic of transmitter plus amplifier and detector can thus be very closely equalized by choosing the proper trans-



WIRING OF AMPLIFIER AND DETECTOR

Figure 3

former to feed the detector, and correct grid impedances for the first tube. A diagram showing amplifier, detector, and equalization is given in Figure 3. The electrical part of the combination can be calibrated experimentally by introducing a small e.m.f. in series with the transmitter and determining the value of this e.m.f. required to give a constant output deflection of the galvanometer. The combination of this characteristic with the sound pressure e.m.f. calibration of the condenser transmitter gives the relation between gal-

[1] Phys. Rev. 10, 39, 1917. 19, 499, 1922.
[2] I. B. Crandall Phys. Rev. 11, 449, 1918.

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vanometer deflection and sound pressure at different frequencies. With proper equalization the maximum deviation from mean in this curve does not amount to more than 5 per cent.

The relation between galvanometer deflection and magnitude of the input potential can be determined in a similar manner. The curve will not in general be a straight line over the whole range so that it is not possible to plot galvanometer deflections as proportional to sound input.

The combination of apparatus described above gives a means for applying a constant potential to a loud speaker at all frequencies, the change in frequency being obtained by turning a single condenser dial. It also includes a sound pickup, amplifier and indicator which will indicate sound pressures at all audible frequencies, correction only being necessary for the non-linearity of the amplitude characteristics. This correction is made by means of the recording apparatus to be described below.

The recording apparatus and oscillator have been designed so as to plot the loud speaker output on paper with a logarithmic frequency scale and linear output scale. The broadness of loud speaker response peaks is in general rather proportional to the frequency. For this reason and also because a more equitable amount of space is accorded to the various regions of the frequency spectrum, logarithmic frequency paper is preferable to linear. It is somewhat more difficult to determine the proper scale to use for intensities. The relation between sound intensity as recorded by the ear and that as given by pressure is a rather complicated function of intensity and frequency. It therefore seems best not to try to represent the ear response but to record sound pressure on a linear scale.

The recording system used is manually operated. That is, a hand operated pointer is made to follow the galvanometer needle and move a pen back and forth on a slowly rotating drum to which the paper is attached. The connection between the pointer and pen (shown in Fig. 4) is built so as to take care of the non-linearity of the detector characteristic, and the motion of the pen is proportional to sound input. The recording drum is geared to the shaft of the condenser which is used to vary the frequency. This condenser has its plates so shaped that when its shaft is rotated with uniform velocity the frequency will change at a logarithmic

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rate. Logarithmic paper can be placed on the recording drum and the plotting will be on a correct scale. The operation of taking a loud speaker curve thus merely involves placing the loud speaker in the correct position before the

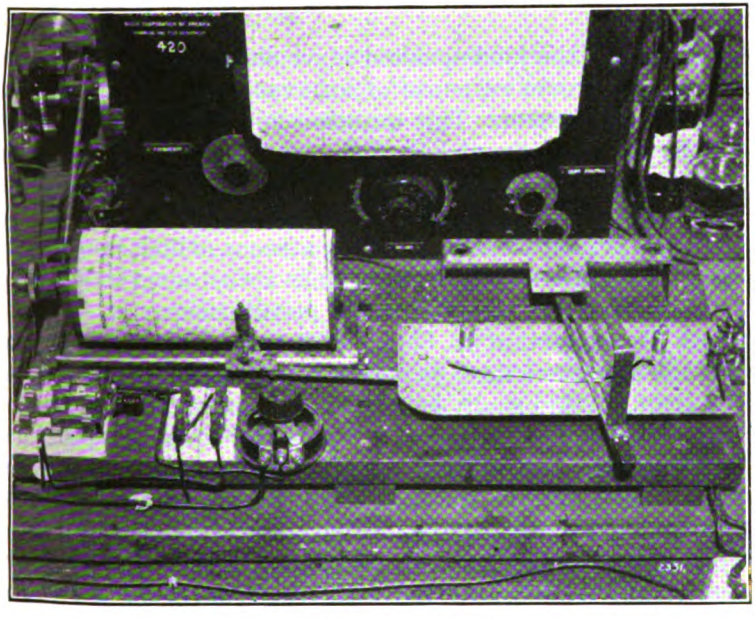

Figure 4—Recording System

microphone, adjusting the output of the oscillator at a predetermined value, and following the galvanometer needle with the pointer while the drum is rotated.

Position of Sound Pickup

The correct position of the microphone with respect to the loud speaker is an important and difficult problem. Offhand it would seem that the logical and simple thing to do would be to place the condenser transmitter about the same distance from the loud speaker as the ear of the listener would ordinarily be, and take a curve. Unfortunately this response is only an approximation to the free wave pressure which would be caused at this place by the loud speaker.

The room in which the loud speaker is placed will greatly affect the curve taken with the microphone placed as described. In the first place, a considerable proportion of the

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sound coming to the microphone gets there indirectly by reflection from the walls and objects in the room. The volume of this sound is naturally affected by the selective absorption and resonance characteristics of the room and our response curve therefore is affected by this room characteristic.

Whether the loud speaker curve should be taken so as to include the room response of an average room is a subject open for discussion. This question is inseparable from that of the microphone pickup and will have to be considered with it. The conclusion will be based on the following premises:

1. A person who has been in a room, which is not average, for a rather long time becomes accustomed to it, and is no longer struck by its acoustic peculiarities.

2. Reproduction to be most natural should in general include the acoustical background of the locale where the program would usually be placed as perceived by the listener in his accustomed position for listening to such a program.

The definition of the perfect reproducing system given at the beginning of the article should then be amplified to say that it should transfer an acoustic condition (pressure and velocity of the sound wave) from the studio to an average room. This will give natural reproduction, for if the room varies from average this will perhaps already have been discounted by the listener.

Our present microphones are essentially pressure measuring devices so that for the present we will have to be content with transferring this quality of the acoustic condition. A little thought will show that this cannot in general perfect reproduction. Assuming then that the microphone plus the amplifier make a perfect transfer of sound pressure to electrical potential, what must the loud speaker do? If the above reasoning is correct, it must in turn make a perfect transfer from electric potential to sound pressure in *an average* room at the *normal listener position.*

There is a point which should be noted regarding the correspondence between the conditions under which the microphone is calibrated and the conditions of its use. The microphones we have been using are calibrated by a method which applies a known pressure of required frequency directly to the diaphragm. (Calibration method described by Wente in article referred to above). The pressure at the

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microphone diaphragm caused by a moving wave is not that present in the wave when no object is in its path. Knowing the size of the microphone and its mechanical characteristics it is possible to get a relation between the true moving wave pressure and the pressure as recorded by the microphone.

At low frequencies the two calibrations will correspond but at the higher frequencies the reflection of the sound wave by the microphone surface will cause the pressure to be twice as great as it would be for the unhampered wave.

When a pure sound is released from some source in a room, a very distinct pattern of nodes and loops of intensity is apparent to a person moving in the room. The microphone having a relatively small surface will be at some point between maximum and minimum intensity. When the frequency is changed the node and loop pattern shifts and in a short range the position of the microphone may be moved from a region of maximum to minimum intensity without any real change in the output of the loud speaker. This gives the curve a series of peaks and depressions which of course obscure the real loud speaker characteristic. This second room effect should not cause an error when averaged over a sufficient frequency range, or region in the room, but may distort the curve very badly at some frequencies, when the microphone is kept in one position.

In order to eliminate this trouble some system of averaging must be developed. One way of doing this would be to rotate or oscillate the loud speaker or microphone or preferably both. At low frequencies the long wavelength and therefore large interference pattern would require a considerable amplitude of motion, comparable with the wavelength. The mechanical difficulties and unwieldiness of this method, in particular when loud speaker in large cabinets are to be tested, make some other way of avoiding this trouble preferable.

As the pickup is moved closer to the loud speaker, a continually smaller proportion of the sound comes by reflection from the walls and objects in the room and the microphone registers more exactly the sound that is coming directly from the loud speaker. Of course when the microphone is too close to the source it acts as a reflecting surface itself and causes an interference pattern to be set up between it and the speaker. The pickup also gets too large

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a proportion of its energy from a limited region of the radiating surface and does not give a true picture of the integrated output. We have found that a characteristic taken at a distance of about fifteen centimeters from the radiating surface gives a quite good indication of the response for loud speakers which radiate in one direction, and in which the radiating surface is not too big.

Loud speakers of the modern cone type however radiate from both sides. When the pickup is placed fifteen centimeters in front of the surface the cone itself will act as sufficient shield to reduce greatly the radiation from the rear which reaches the microphone. In particular when the loud speaker is placed in a cabinet the long path around will

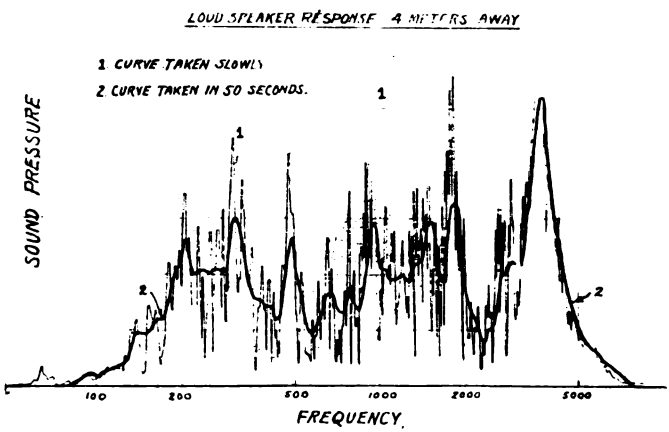

Figure 5

cause so much attenuation as to practically eliminate this part of the radiation. In loud speakers placed in cabinets the sound off the rear may be very important, as cabinet resonance will usually show up as radiation in this direction. Of course the obvious thing to do in a case like this is to take a second curve from the rear. The actual response will be some combination of the two curves, depending on the position of the listener.

There is still one serious error made in adding together the response of loud speakers taken at this small distance from front and rear. Sound is a wave disturbance and in any addition phase differences must be taken into account. This becomes particularly important at low frequencies. The radiation from the rear is naturally 180 deg. out of phase

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with that from the front, from single vibrating surfaces. If the path lengths to the listener differ by only a fraction of a wave, destructive interference will be set up. It then seems as if some kind of curve at a distance is still necessary. Since the characteristics taken at fifteen centimeters will show the fine structure of the loud speaker spectrum, such as sharp peaks and depressions, a rather rough curve at a distance will be sufficient.

If the galvanometer used to record the response does not have an extremely short period it is possible to change the frequency quickly enough so that the needle will not have time to respond to the room peaks and at the same time will be able to record the gross loud speaker characteristics.

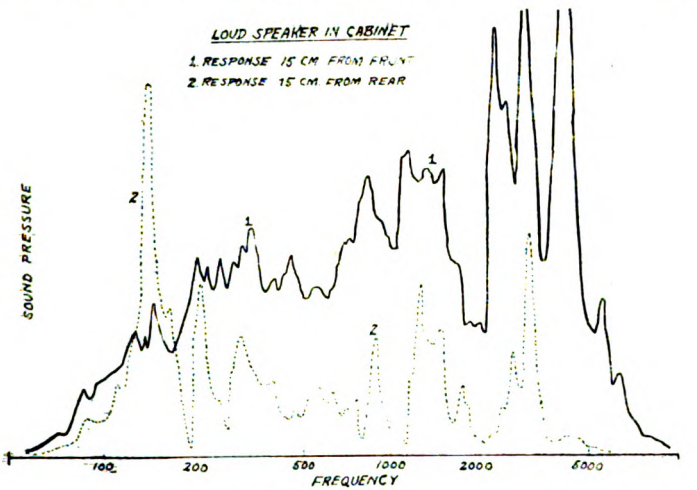

Figure 6-a

This curve we find gives a very good representation of the general response. Fig. 5 gives a comparison of a curve taken slowly at a distance with one taken as described.

Summary

The procedure of making a loud speaker test will be made a little clearer if we summarize here.

1. The output of an oscillator, continuously variable, and supplying constant potential, is fed to the loud speaker through a resistance equal to the plate impedance of the tube to be used with the loud speaker. Three curves showing sound output are taken as follows:

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(a) Microphone fifteen centimeters in front of loud speaker, frequency varied slowly enough to show all peaks.

(b) Same with microphone fifteen centimeters to the rear.

(c) Microphone at normal listener's distance from loud speaker in an average room; oscillator condenser driven by motor quickly enough so that room peaks will not be recorded and slowly enough to show principal loud speaker characteristics.

Figure 6 shows three curves taken on the same loud speaker by methods (a), (b) and (c). The above test may

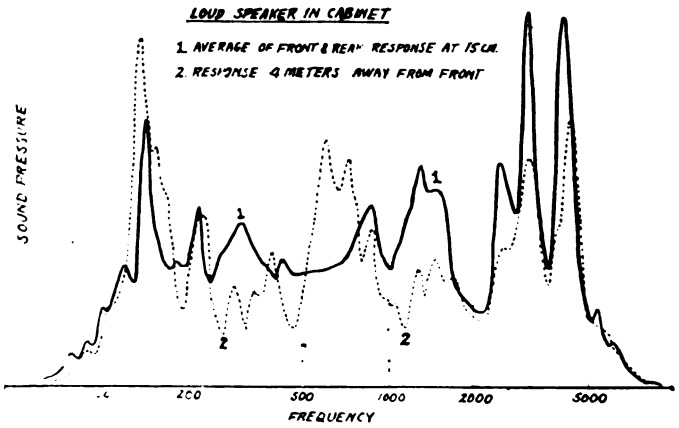

Figure 6-b

seem needlessly lengthy, but the time involved is really not great. The first two curves require about three to four minutes for recording and the last one takes rather less than a minute.

2. We have found that the loud speaker which has the best looking characteristic (most free from peaks) will in spite of the defects of existing broadcasting transmitters and receivers generally sound best when tried on radio.

We have never built a perfect loud speaker however, and the question always arises as to the relative undesirability of defects which can be substituted for each other. The only way to determine this is to try the loud speaker on a set receiving normal broadcasting.

When the loud speaker is to be used exclusively for phonograph reproduction, a phonograph test is of course

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substituted for the one with radio. When it is to be used in combination equipment both tests are required and a compromise must be reached, as the phonograph recording characteristic is, in spite of recent improvements, still different from the radio set response.

We have also at times gone somewhat farther in the test of the combination of set with loud speaker, using the oscillator to modulate a broadcast frequency wave which is impressed on the set. The mechanics of taking the curve is the same as that described when the oscillator is im-

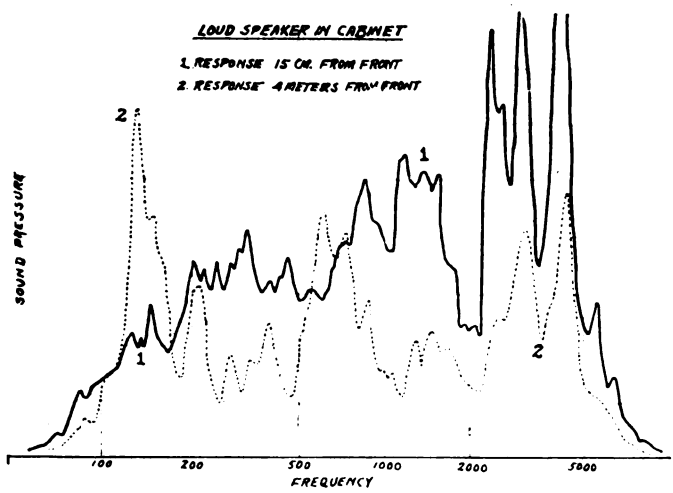

Figure 6-c

pressed directly on the loud speaker. The curve taken in this manner gives the response of the set plus loud speaker for uniform modulation of a constant amplitude broadcast frequency wave over the audible range.

The results of such tests are very interesting and show quite clearly why a poor quality loud speaker may at times sound better on a receiving set than a speaker which measures up well. It is very true however that the loud speaker which shows up best on the original tests described above seldom fails to sound best with a receiving set.

We are looking forward to the time when we may have the ideal condition when the oscillator can be impressed on the input end of the broadcast studio and a radio set tuned to this station will give an output through its loud speaker

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which draws a straight horizontal line across our curve sheet.

In concluding we would like to mention that Dr. John P. Minton was associated with us in a part of the work outlined above, and some of the ideas given are due in no small measure to this association.

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HIGH ANGLE RADIATION OF SHORT ELECTRIC WAVES *

BY
S. UDA

(Tohoku Imperial University, Sendai, Japan.)

(I) INTRODUCTION

As a means of producing high angle radiation of short radio waves, it is usual to make a straight vertical antenna operate at one of its harmonics. The field distribution of such an antenna has been theoretically calculated by B. van der Pol, S. Ballantine, S. A. Levin and C. J. Young, and others. No experimental work confirming these theoretical results seems to have ever been carried out.

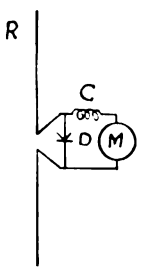

Figure 1

The present article contains some accounts of experimental work in connection with this, and also gives the test results on a new wave projector devised by the author with special reference to high angle radiation of short electric waves.

(II) APPARATUS USED

The short wave generator used by the author is of a push-pull type which can produce stable continuous waves

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of few meters, in length. The receiving circuit is shown in Figure 1, where R is the receiving antenna forming a Hertz's resonator, and D and M are a crystal detector and a micro-ammeter respectively.

Now the receiving system is mounted in a wooden case and moved up and down along a vertical line by a pulley device as shown in Figure 2. A and B are two wooden poles erected 3.5 m. apart, the height of which being nearly equal to 10 meters.

The generator is placed at a certain distance from the

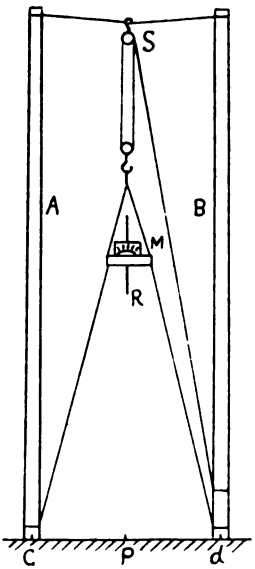

Figure 2

receiver and set close to the ground level. Field intensity due to the generator can thus be measured in various angular directions above the ground level by placing the receiver at various heights. The meter reading must be taken from a distant point through a telescope, for the observer's body causes a remarkable effect on the field distribution.

(III) Field Due to a Straight Vertical Antenna

In order to facilitate the change of antenna length, a straight telescopic brass tube was employed as the send-

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ing antenna. The antenna was unloaded and the operating wave length was kept constant at 2.66 meters throughout the experiments.

Since the wave length was kept constant, the total

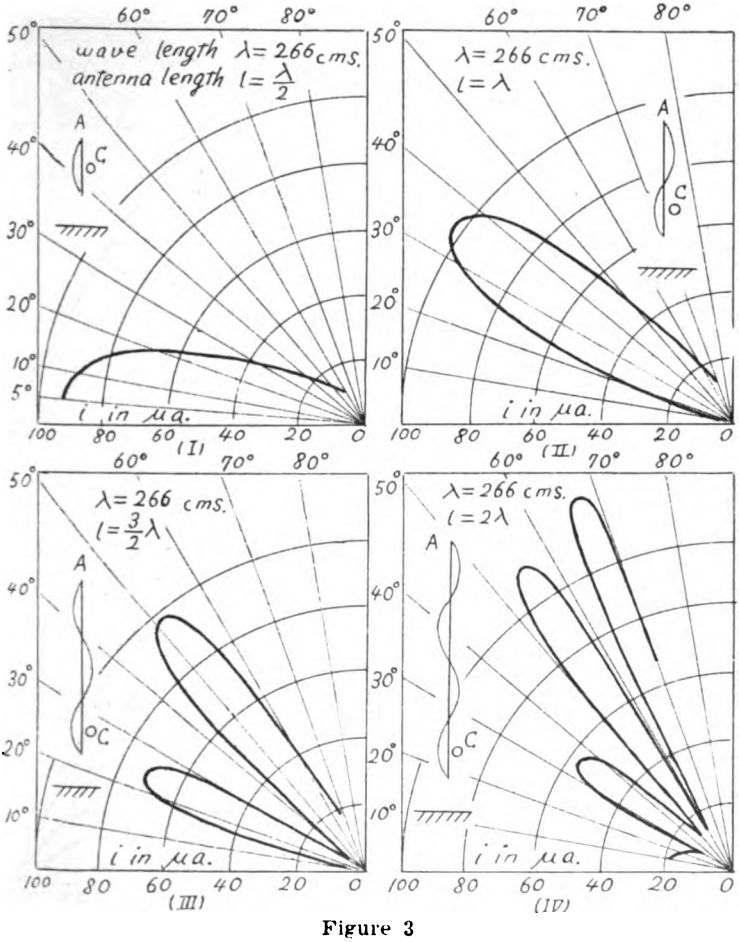

Figure 3

length of the sending antenna had to be altered according to the harmonic adopted. The grounded and the ungrounded antennae were used in the experiments.

(a) *Grounded Antenna.* When the lower end of the antenna is grounded, oscillation is possible at odd harmonics only, whereas the ungrounded antenna can operate either at an even or an odd harmonic.

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Figure 3 shows the observed polar curves in the case of the grounded antenna, and the radius vector gives the measure of intensity in the receiving system placed in that direction. *A* is the sending antenna, and the current dis-

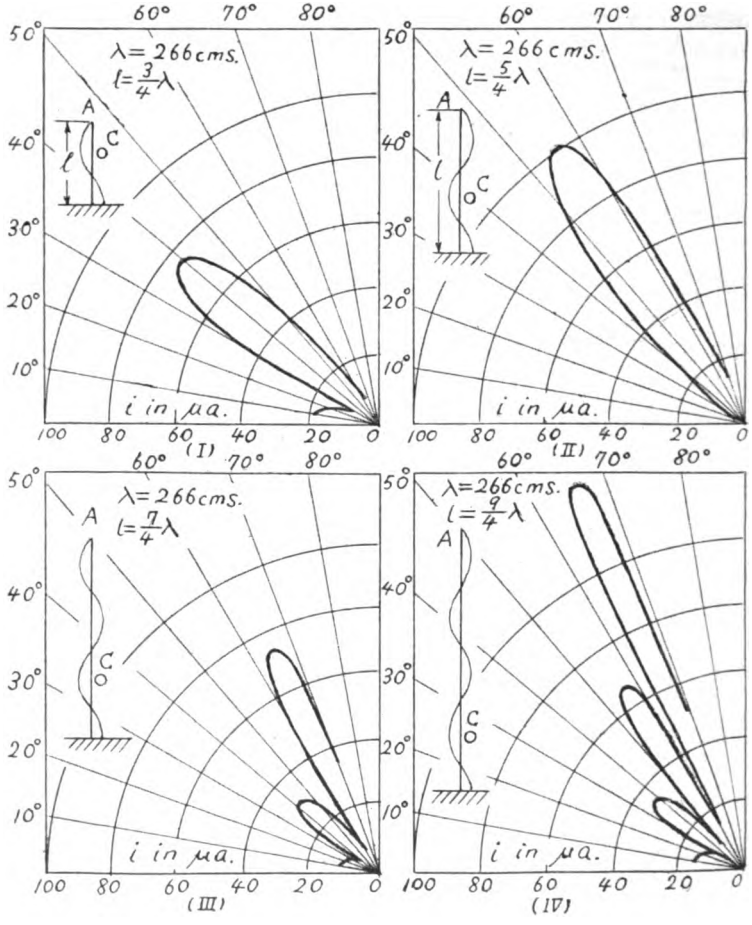

Figure 4

tribution in it is assumed to be a sine curve as indicated (Figure 3). *C* represents the oscillating coil of the short wave generator which is coupled with the antenna and placed near its current loop.

In all cases, it will be noticed that a considerable amount of energy is radiated in high angular directions,

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while the radiation along the earth surface is very small. If the length of the antenna is made equal to $\frac{7}{4}\lambda$ or $\frac{9}{4}\lambda$ corresponding to the 7th or the 9th harmonic, the wave projected upwards will be split up into several parts and there will then be several maxima of radiation. This is illustrated by (II) and (IV) in Figure 3.

It must be remembered that in our experiments only the vertical component of the field has been measured and

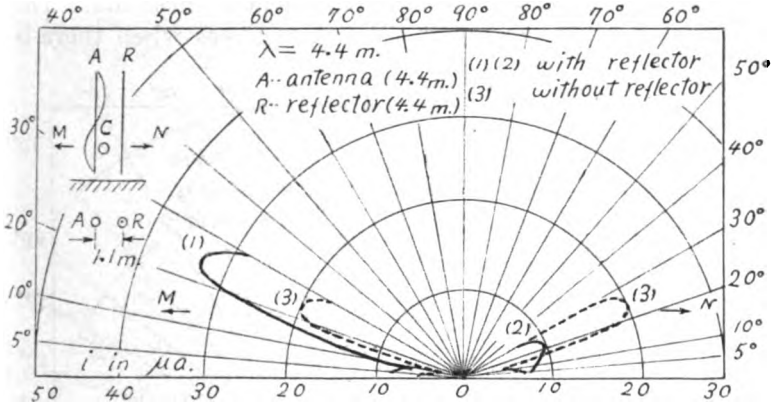

Figure 5

therefore the polar curves can not be strictly compared with the theoretically determined curves.

(b) *Ungrounded Antenna.* The same measurements were made with the ungrounded antenna. In this case, the lower end of the antenna was kept at a distance of about a quarter wave length above the ground. The observed polar diagrams are shown in Figure 4, where the horizontal base line is taken along the earth surface.

Here again is it noticeable that the radiation at higher angles is markedly predominant when the antenna oscillates at one of its harmonics.

(IV) High Angle Directional Beam

Suppose that a vertical antenna operating at the 2nd harmonic is sending out electromagnetic wave in all directions around it. If a single metallic rod of a full wave length is vertically erected at a distance of a quarter wave

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length behind the radiating antenna (Fig. 5), then this metallic rod, as is well known, will act as a wave reflector. Since this wave reflector will also act at the 2nd harmonic, it is obvious that the characteristic of the high angle radiation will become much augmented in a single direction due to the existence of the reflector. Thus we can easily obtain a high angle directional beam of short radio waves.

In Figure 5 is shown the directive effect of a single wave reflector. The curve (1) shows the radiating characteristic in the forward direction of the antenna, and (2) represents the same in the backward direction. The broken lines (3) represent the polar curves when there is

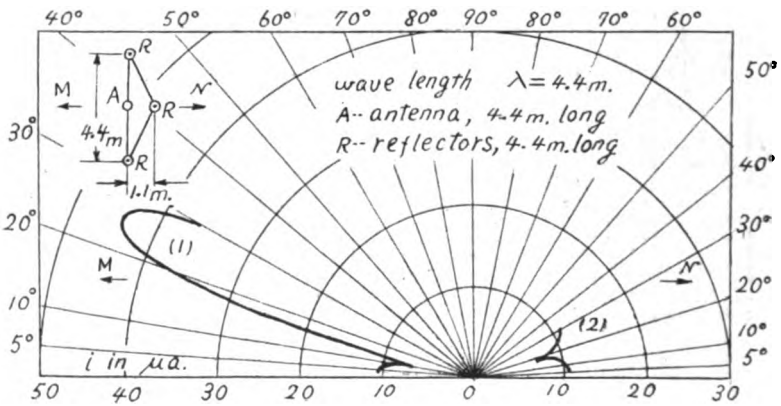

Figure 6

no reflector. The operating wave length is here equal to 4.40 meters, and both the antenna and the reflector are ungrounded, the lower ends of them being at a height of a few centimeters from the ground.

Again a reflecting rod is placed a quarter wave behind the antenna and two more reflectors, one being on the left and the other on the right side of it, are placed a half wave distant from the antenna (Fig. 6). These three rods form a triple-antenna reflecting system, which is comparatively simple in form yet fairly effective in action. For the future convenience, this reflecting system will hereafter be called a fundamental "trigonal reflector."

It is, of course, most efficient when these reflectors are all tuned to the 2nd harmonic, whereby they have nearly the same length as the main antenna. Figure 6 gives the

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observed diagrams in this case, where (1) and (2) represent the curves in front and in back of the antenna respectively.

In Figure 7 is shown the radiating characteristic in the side directions under the same operating conditions as in Figure 6.

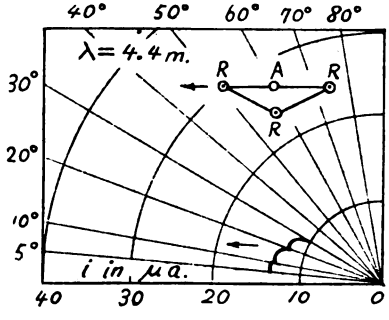

Figure 7

Now if several wave directors are arranged in front of the radiating antenna along the line of reflection, the directivity will be remarkably improved. This is evident from Fig. 8 and Fig. 9, for example, where six wave directors are employed.

The *D's* in the Figure are the directors which are hung along an inclined line lying nearly in the direction of maximum radiation (Fig. 8.) The length of each director is

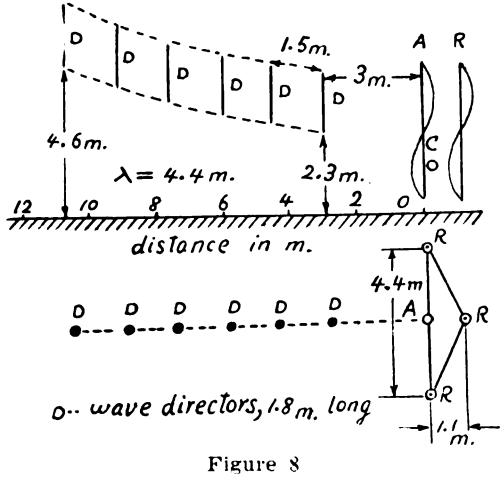

Figure 8

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1.80 meters, and their separation 1.50 meters. Figure 9 shows the observed polar distribution in this case. For all the observations of Fig. 5 and Fig. 9, the field measurements were made under the same condition, and the short wave generator was also kept at exactly the same condition. The distance between the generator and the receiving apparatus was also made the same for all the above cases and was equal to 15 meters.

The series of wave directors acts as a means of converging the wave energy and transmitting it further along its direction, and may therefore, be called a "wave duct" or a "wave canal." The projection of the sharpest beam

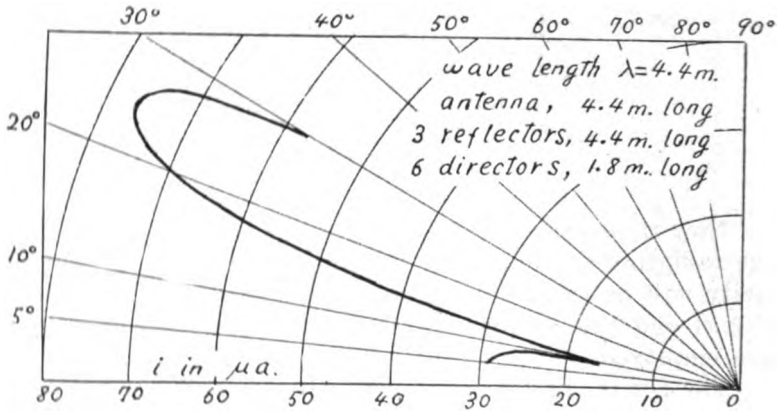

Figure 9

can be effected by combining a trigonal reflector with a wave canal. This combination is thus called a "wave projector." The sharpness of the beam can be improved by increasing the number of director rods contained in the wave canal.

Many observations have been made by the author on other types of wave projector, in which the antenna and reflectors operate resonantly at a half wave length. Polar curves thus obtained have proved that a beam with the sharpness never before attained could then be produced.

As regards the general study of this wave projecting system, a preliminary report was presented to the Imperial Academy of Japan. (YAGI and UDA, "Projector of the sharpest beam of electromagnetic waves," Proceedings Imp. Academy 2, 1926.) The full account of the study

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in detail has been published as a series of papers in the Journals of I. E. E., of Japan. (UDA. On the wireless beam of short elec. waves. I to VII written in Japanese). Two papers relating to the same subject were presented to the 3rd Pan-Pacific Science Congress held in Tokyo, November, 1926. (YAGI and UDA. A new electromagnetic wave projector and radio beacon. YAGI and UDA. On the feasibility of power transmission by electromagnetic waves).

In conclusion, the writer wishes to acknowledge his indebtedness to Saito Memorial Foundation for the grant which enabled him to undertake the work, and to Professor H. Yagi, of the Tohoku Imperial University, under whose direction the work was carried out.

SUMMARY

The paper describes some accounts of experimental work on the field distribution due to a straight vertical unloaded antenna operating at one of its harmonics. Short wave of 2.66 meters length were employed, and observations have been made with the grounded and the ungrounded antennas.

The paper also gives the test results on a new wave projector devised by the author with special reference to high angle radiation of short electric waves.

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NOTES ON RADIO RECEIVER MEASUREMENTS*

By

THEODORE A. SMITH AND GEORGE RODWIN

(Technical and Test Department Radio Corporation of America)

It is very desirable to have a basis of comparison for radio receivers of different types. Methods of test for commercial electrical apparatus have been well established, and test codes for motors, boilers, engines and similar equipment have been adopted by engineering societies. At the present time, however, the specifications of the performance of a radio set are still in a more or less rudimentary stage. The purpose of this paper is to present some contributions to this subject.

Methods of test which enable diverse types of receivers to be compared and which would be of practical application were suggested some time ago by Mr. Julius Weinberger, of this department. Tests have been made on various receivers and the results obtained seem to warrant a description of the methods used.

In order to compare radio receivers electrically, it is necessary to select certain essential attributes which determine their performance. Different sets can then be compared with respect to these quantities. Three characteristics have been chosen as most important for measuring the performance of a receiver. They are Sensitivity, Selectivity and Fidelity. Two other factors of slightly lesser importance are the Accepted Frequency Range and the ability to control the volume of the output.

The essential characteristics mentioned above may best be shown by a series of curves. Sensitivity may be plotted as the receiver output at various radio frequencies. Selectivity may be indicated by a Resonance Curve. Fidelity can be shown in two ways. In the first, an overall curve may be plotted of the audio frequency output when constant amplitude modulation at progressive audio frequen-

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cies is received. While this curve represents the quality which may be expected, it is impossible to tell to what part of the set a deviation is due. A better method is to take a curve on the radio circuits, which shows the sideband transmission characteristics, and one of the audio amplifier characteristics. A combination curve may then be obtained, from all frequency selective circuits of the receiver, which will represent the quality to be expected from the set.

Method of Test

The attempt has been made, in testing receivers, to simulate actual conditions of normal use. The function of a radio set is to produce a definite loud speaker signal, and the set should be rated upon its ability to do this. A sensitive receiver is one which requires relatively little voltage from the antenna to actuate the loud speaker. A perfectly selective receiver requires an infinite field strength from an interfering station, to produce an audible loud speaker signal.

In order to determine the various factors mentioned previously, it is first necessary to define what is meant by a "loud speaker signal." The term "standard loud speaker signal" or simply "standard signal" has been taken to correspond as nearly as possible with the normal loud speaker output of a receiver as used in the home. This may be regarded as a highly indefinite term, but it is for this highly indefinite condition that receivers are being built, and as there is no ready-made term to define such a condition, it has been necessary to manufacture one. The "standard loud speaker signal" must be great enough to enable persons a distance away from a loud speaker, to hear a program distinctly and without effort. It must also be limited by the output capacity of the final amplifier tube. The standard loud speaker signal employed, has been arbitrarily defined as the signal corresponding to an average audio frequency (r. m. s.) voltage of 15 volts across a 5000 ohm resistance in the output circuit when 400 cycle modulation at 50% is furnished by the signal generator.

While such a "standard signal" may be useful to talk about and to form a basis of comparison, it is not easy to

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measure in practice. For this reason it becomes necessary to select an electrical analogue for measurement purposes. This is not difficult, as the detector of the radio set may be employed as an indicator by placing a milliammeter in its plate circuit. If a signal of constant degree modulation is being received, a certain change in detector plate current will correspond exactly to the above mentioned "standard signal." This value of plate current, or change in plate current from static value, will be used as a measure of the standard signal throughout. It is then unnecessary to use modulated signals for receiver measurements, although the amplification of the audio stages is taken into account.

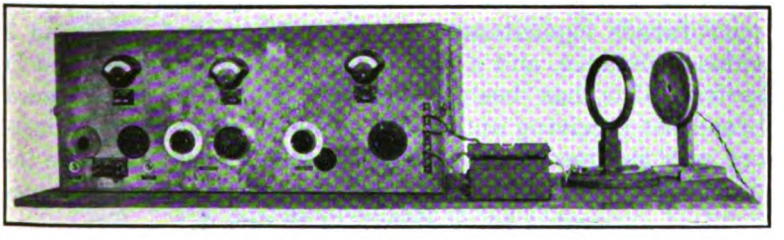

Figure 1—Test Oscillator used for taking all data. System comprises an audio frequency amplifier, modulators, oscillator, and power amplifiers with output metering device and modulation percentage indicator.

The manner in which the given loud speaker voltage is related to the detector plate current change, may be determined by experimental means or else by calculation. To compute it, the gain of the audio amplifier must be known. The loud speaker voltage, 15 volts, divided by this value will give the audio frequency voltage in the plate circuit of the detector. If an average degree of modulation of 50% is assumed, this audio frequency voltage will bear a definite relation to the voltage of the carrier frequency on the input side of the detector, and will be accompanied by a definite plate current change (dc) of the detector. This current value is a measure of the "standard signal."

All other factors are related to this value. Sensitivity curves may be taken as the field strength required to produce the "standard signal" or standard detector plate cur-

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rent change, at various radio frequencies, when the receiver is tuned to resonance at each frequency.

Resonance curves represent the field strength necessary to produce the standard signal at various radio frequencies when the receiver is left tuned to resonance at one particular frequency.

A volume control curve can be plotted of field strength required to produce standard signal when the receiver is tuned to resonance, versus settings of the volume control.

Method of Taking Curves

In all cases a radio frequency oscillator is necessary. For coupling to receivers requiring an antenna, a dummy antenna was provided consisting of a condenser of 0.0004 microfarads capacitance, a coil of 28 microhenrys and 2 ohms resistance, and a resistance of 23 ohms, connected in series. This has been found to simulate a fairly good broadcast receiving antenna. An effective height of 4 meters was assumed. The oscillator coil is coupled to the inductance coil of the dummy antenna, the general arrangement of apparatus being shown in Figure 1.

If the oscillator coil has a current I flowing in it at some radio frequency f and there is a mutual inductance M between it and the inductance in the dummy antenna, a radio frequency voltage $2\pi fMI$ will be induced in the dummy antenna circuit. This voltage could be induced by a field strength E in an equivalent actual antenna of effective height H, so that $E=\frac{2\pi fMI}{H}$ In using this expression E is in volts per meter, f in cycles per second, M in henrys, and I in effective amperes.

For loop receivers, the oscillator coil is used to produce known field intensities at the loop of the set. A coil of n turns with radius a and having a current I flowing in it will produce a flux density H at a distance x given by

$$H=\frac{2\pi a^2 In}{10(x^2+a^2)^{3/2}} \qquad \textit{(gausses)}$$

Expressed as equivalent microvolts per meter this becomes

$$E=300H\times 10^6=18850\frac{na^2 i}{(x^2+a)^{3/2}}\cos\alpha \cdot 10^6$$

in which all distances are expressed in centimeters, i in am-

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peres, α is the angle taking the planes of the two coils.

For taking sensitivity curves, the receiver under test is tuned to resonance, the oscillator output is adjusted so that the detector plate current change reaches the proper value for standard signal. Several points at different frequencies may be taken quickly in this manner and the sensitivity curve plotted as the field strength required to produce standard signal, versus frequency.

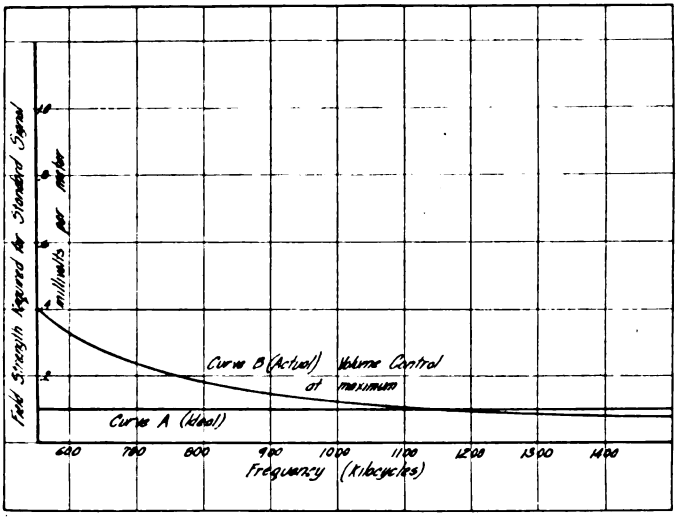

Figure 2—Sensitivity Curve—Field Strength Required to Produce Standard Signal vs. Frequency.

Resonance curves may be taken in two ways. By the first method, an oscillator is necessary in which the frequency can be varied in small steps and the output of which can be varied widely without change in frequency. A neutralized master oscillator-power amplifier system has been used. (See Fig. 1). The receiver is tuned to resonance and the oscillator adjusted to this frequency. The oscillator frequency is then varied and readings of field strength are taken at several frequency steps off tune, keeping the detector plate current change constant at the standard signal value.

In case such an oscillator is not available, an oscillator the output of which is constant over a small frequency range may be employed. In this method, a resonance

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curve is taken by keeping the oscillator output constant and reading the detector plate current changes for steps of frequency off resonance. By then plotting a curve of field strength versus plate current changes, this resonance curve may be translated to the previous form.

The volume control curve is self-explanatory.

From the resonance curve is shown the effect of the

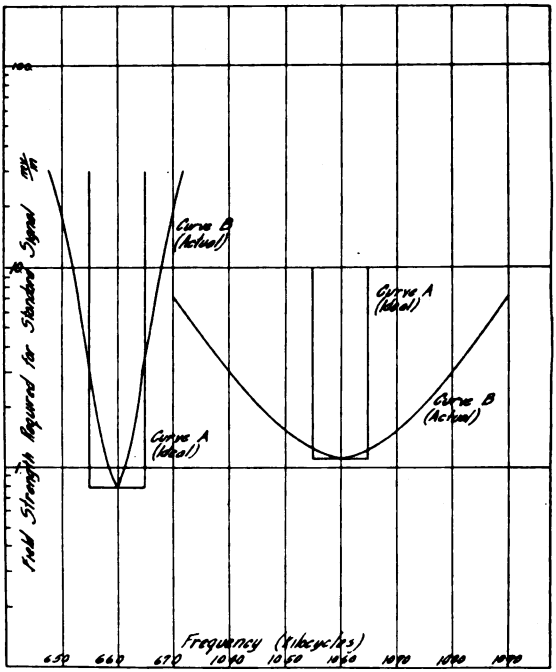

Figure 3—"Resonance" Curves—Curve at left taken with receiver tuned to 660 kc., curve at right with receiver tuned to 1060 kc.—Field Strength Required to Produce Standard Output vs. Frequency of Signal Generator.

receiver selectivity in cutting off the higher audio frequencies. However, it is more convenient to show this sharpness by replotting a portion of the resonance curve in different form. This new curve shows in effect the audio frequency voltage transmitted by the radio frequency circuits, plotted as a percentage of the transmission at the resonant frequency. As this quantity is proportional to the admittance of the circuits, the title of the ordinate, Percent. Resonant Admittance, seems a proper one. The

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data for this curve are obtainable by taking the average of the field strengths required to produce standard signal either side of resonance for frequencies in the audio

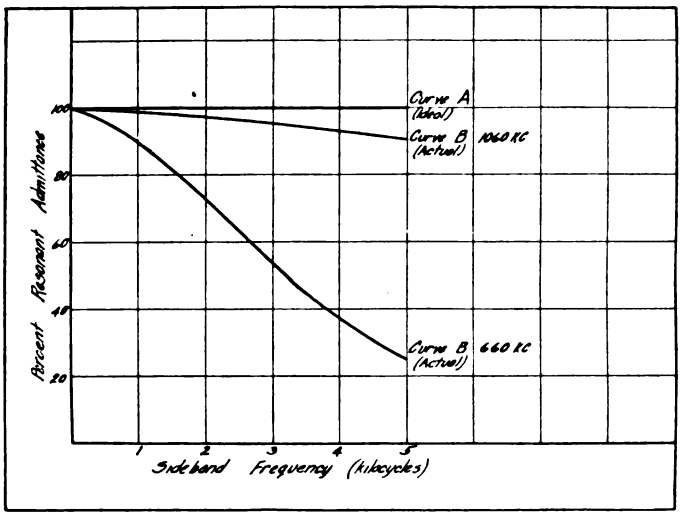

Figure 4—Sideband Transmission Characteristic Curves—Cutoff of high audio frequencies due to selectivity of radio frequency circuits.

range. This average field strength divided into the field strength required to produce standard signal at resonance, is the Resonant Admittance expressed as a percent.

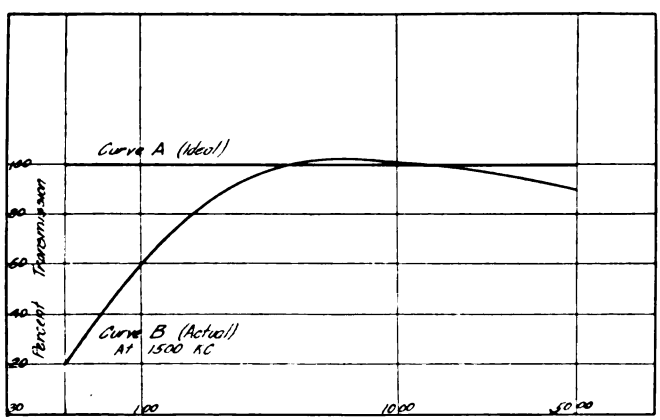

Figure 5—Audio Frequency Characteristics of Entire Receiver—Transmission of audio frequencies compared to transmission at 400 cycles.

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By combining this curve with the curve of the audio amplifier, the frequency response curve of the receiver will be obtained, if there is no other frequency selective circuit, (such as the grid leak condenser combination). As the latter effect can be allowed for, the combined curve may be taken as truly representing the quality limitations of the receiver. As measurements are taken below the overload-

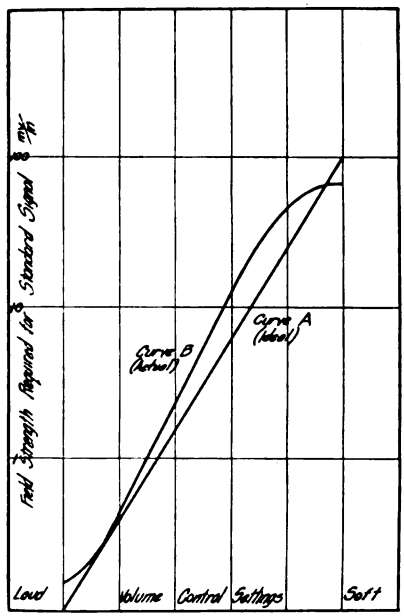

Figure 6—Volume Control Curve—Receiver and Signal generator in tune. Field strength to produce standard output at various volume control settings.

ing point of the amplifier tubes, no serious harmonic distortion is encountered.

The method of procedure for quickly taking audio frequency characteristic curves will be described in an accompanying paper by Mr. E. T. Dickey.

While the description of the method of test makes the process appear involved, the actual operation is relatively simple and curves may be taken with considerable speed.

Relation of Actual to Ideal Curves

When the curves of a receiver have been obtained, they may be compared with curves of other sets. Perhaps a

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good method of comparison is to note the deviation from a set of hypothetical, ideal curves which represent the ultimate aim of receiver design.

The shape of a series of ideal curves is shown in *A* of Figures 2, 3, 4, 5, and 6. Curves marked *B* in these figures represent actual curves of some receivers tested.

CONCLUSION

While it is not claimed that the methods employed herein are new, few attempts have been made previously to coordinate the testing of receivers, so that the results obtained may be correlated with broadcasting station performance. It is possible to predict how well a set in a given locality will receive signals from a broadcasting station, by inspection of the sensitivity curve of the set and a field strength map of the station. The quality which may be expected from a receiver is found from the frequency characteristic curves. The field strength which still overloads the set with the volume control device at its minimum position can be determined, and hence the area about a broadcasting station in which such a receiver cannot be used, is known. It is possible to tell from a resonance curve whether a broadcasting station will interfere with one being received if their respective field intensities are known. Thus it will be seen that the results obtained in this method of test are useful for practical as well as theoretical purposes.

It is to be hoped that some general method of test will be adopted, so that in the future receivers may be compared by means of definite engineering standards.

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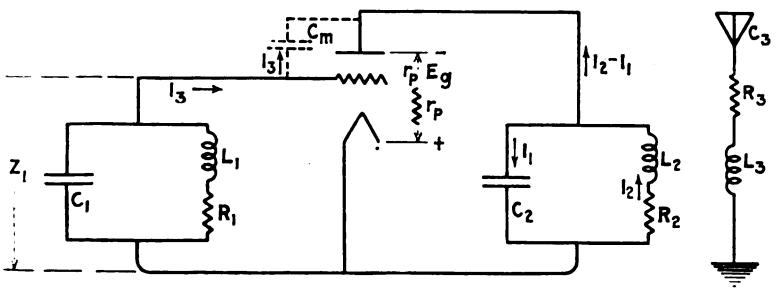
THE TUNED-GRID TUNED-PLATE CIRCUIT USING PLATE-GRID CAPACITY FOR FREED-BACK. A DERIVATION OF THE CONDITIONS FOR OSCILLATION.

By
J. B. Dow

Herewith is a circuit which has recently been popularized in short wave transmission. The necessary feedback for oscillation generation is obtained solely through the plate-grid capacity of the tube and not through external inductive or capacity coupling as in most other circuits.



The grid-filament and plate-filament capacities within the tube are considered to be lumped with the tuned circuit capacities C and C_2 respectively. R_1 is the equivalent resistance of the input resistance of the tube plus such resistance as is inherent in the tuned grid circuit. R_2 is the equivalent resistance of the load circuit L_3, R_3 C_3 plus the inherent resistance of the tuned plate circuit. L_2 and C_2 include the equivalent inductance and capacity of the load circuit L_3, R_3 C_3.

The problem is to develop equations showing the conditions required for oscillation.

It is assumed that the alternating current wave is sinusoidal when oscillation occurs. Owing to the presence of the

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tuned grid circuit, this assumption introduces no serious error in the results.

Kirchoffs Laws give

$$I_1\frac{j}{C_2\omega}-I_2(R_2+jL_2\omega)=0 \tag{1}$$

$$(I_2-I_1)r_p+I_1\frac{j}{C_2\omega}=\mu E_g \tag{2}$$

Note: All *I's*, *E's* and *z's* are complex.

Digression on E_g:

$$\frac{1}{z_1}=\frac{1}{R_1+jL_1\omega}-\frac{1}{\frac{j}{C_1\omega}}$$

$$z_1=-\frac{(R_1+jL_1\omega)\frac{j}{C_1\omega}}{j(L_1\omega-\frac{1}{C_1\omega})+R_1}$$

$$I_3=\frac{-I_1\frac{j}{C_2\omega}}{z_1-\frac{j}{C_m\omega}}=\frac{I_1C_1\omega C_m\omega[R_1+j(L_1\omega-\frac{1}{C_1\omega})]}{C_2\omega[[C_m\omega(R_1+jL_1\omega)+C_1\omega[R_1+j(L_1\omega-\frac{1}{C_1\omega})]]]}$$

$$E_g=I_3z_1=-\frac{-jI_1C_m\omega(R_1+jL_1\omega)}{C_2\omega[[C_m\omega(R_1+jL_1\omega)+C_1\omega[R_1+j(L_1\omega-\frac{1}{C_1\omega})]]]} \tag{3}$$

Substituting (3) in (2), get

$$I_1[\frac{j\mu C_m\omega(R_1+jL_1\omega)}{C_2\omega[[C_m\omega(R_1+jL_1\omega)+C_1\omega[R_1+j(L_1\omega-\frac{1}{C_1\omega})]]]}+\frac{j}{C_2\omega}-r_p]+I_2r_p=0 \tag{4}$$

$$I_1\frac{j}{C_2\omega}-I_2(R_2+jL_2\omega)=0 \tag{1}$$

Eliminating between (4) and (1) and multiplying both sides of the result by $C_2\omega$, get

$$jr_p+[\frac{j\mu C_m\omega(R_1+jL_1\omega)}{[[C_m\omega(R_1+jL_1\omega)+C_1\omega[R_1+j(L_1\omega-\frac{1}{C_1\omega})]]]}+j-r_pC_2\omega](R_2+jL_2\omega)=0 \tag{5}$$

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Rationalizing the denominator of the fraction, then multiplying both sides of (5) by the new denominator, get

$$jr_p[R_1^2(C_m\omega+C_1\omega)^2+[[C_m\omega L_1\omega+C_1\omega(L_1\omega-\frac{1}{C_1\omega})]]^2]$$
$$+\Big\{j\mu C_m\omega(R_1+jL_1\omega)[[R_1(C_m\omega+C_1\omega)-j[C_m\omega L_1\omega$$
$$+C_1\omega(L_1\omega-\frac{1}{C_1\omega})]]]$$
$$+j[[R_1^2(C_m\omega+C_1\omega)^2+[C_m\omega L_1\omega+C_1\omega(L_1\omega-\frac{1}{C_1\omega}]^2]]$$
$$-r_pC_2\omega[[R_1^2(C_m\omega+C_1\omega)^2+[C_m\omega L_1\omega+C_1\omega(L_1\omega-\frac{1}{C_1\omega}]^2]]\Big\}$$
$$(R_2+jL_2\omega)=0 \qquad (6)$$

Equation (6) expresses in complex form the conditions required for oscillation. The real part of (6) gives the first general condition of oscillation, viz:

$$\mu C_m\omega R_1(C_m\omega+C_1\omega)(R_2L_1\omega+R_1L_2\omega)+[C_m\omega L_1\omega+C_1\omega(L_1\omega-\frac{1}{C_1\omega})](L_1\omega L_2\omega-R_1R_2)]$$
$$-[[R_1^2(C_m\omega+C_1\omega)^2+[C_m\omega L_1\omega+C_1\omega(L_1\omega-\frac{1}{C_1\omega})]^2]](L_2\omega+R_2r_pC_2\omega)=0 \qquad (7)$$

If the tuned grid circuit is adjusted to resonance, $L_1\omega=\frac{1}{C_1\omega}$, then the above equation reduces to the simpler form

$$\mu C_m\omega[R_1(C_m\omega+C_1\omega)(R_2L_1\omega+R_1L_2\omega)+C_m\omega L_1\omega(L_1\omega L_2\omega-R_1R_2)]$$
$$-[R_1^2(C_m\omega+C_1\omega)^2+(C_m\omega L_1\omega)^2](L_2\omega+R_2r_pC_2\omega)=0 \qquad (8)$$

If $L_1\omega=\frac{1}{C_1\omega}$, and in addition, $R_1=0$, (7) reduces to the still simpler form

$$\frac{L_2}{C_2}=\frac{R_2r_p}{\mu-1} \qquad (9)$$

The imaginary part of (6) gives another condition for oscillation from which the frequency can be determined. This second general condition is

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$$\mu C_{m\omega}[[R_1(C_{m\omega}+C_{1\omega})(R_1R_2-L_{1\omega}L_{2\omega})+[C_{m\omega}L_{1\omega}+C_{1\omega}(L_{1\omega}-\frac{1}{C_{1\omega}})](R_2L_{1\omega}+R_1L_{2\omega})]]$$
$$+[[R_1^2(C_{m\omega}+C_{1\omega})^2+[C_{m\omega}L_{1\omega}+C_{1\omega}(L_{1\omega}-\frac{1}{C_{1\omega}}]^2]][R_2+r_p(1-L_{2\omega}C_{2\omega})]=0 \quad (10)$$

If as before, $L_{1\omega}=\frac{1}{C_{1\omega}}$, (10) reduces to

$$\mu C_{m\omega}[R_1(C_{m\omega}+C_{1\omega})(R_1R_2-L_{1\omega}L_{2\omega})+C_{m\omega}L_{1\omega}(R_2L_{1\omega}+R_1L_{2\omega})]$$
$$+[R_1^2(C_{m\omega}+C_{1\omega})^2+(C_{m\omega}L_{1\omega})^2][R_2+r_p(1-L_{2\omega}C_{2\omega})]=0 \quad (11)$$

If $L_{1\omega}=\frac{1}{C_{1\omega}}$, and in addition, $R_1=0$, (10) reduces to

$$R_2(\mu-1)+r_p(1-L_{2\omega}C_{2\omega})=0 \quad (12)$$

from which

$$\omega=\sqrt{1+\frac{R_2(\mu+1)}{r_p}}\cdot\frac{1}{\sqrt{L_2C_2}} \quad (13)$$

The last equation indicates that if $R_2\neq 0$, the frequency of oscillation is greater than the undamped frequency of the circuit $L_2C_2R_2$.

$\omega=2\pi f$.

μ=plate amp. factor.

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SELECTIVITY OF TUNED RADIO RECEIVING SETS*

KENNETH W. JARVIS

(Engineer, Crosley Radio Corporation)

In the design of modern broadcasting receivers, the radio engineer has three technical objectives. These three objects are the obtaining of selectivity, fidelity of reproduction and adequate sound volume. Selectivity means absolute choice of the wanted transmission, without interference of any kind. Adjacent stations, static and man-made interference should be entirely eliminated. Fidelity of reproduction defines itself. The final sound output should be exactly similar in amplitude, phase and tone as that before the transmitter microphone. Adequate sound volume implies sufficient amplification to increase any signal impulse, no matter how small, to any desired value.

An ideal radio receiver will have all of these factors incorporated in its design. The user may choose his program and control the output to suit his comfort. These should be independent problems, a more perfect solution of any of them contributing to the intrinsic value of the receiver. Unfortunately this situation is not even approached in present practice—a change in design which will affect any one of these factors will also affect the other two. It will therefore be appreciated that a paper on "Selectivity" would be meaningless if the entire combination of effects were not considered. The most that can be done is to show how the amplification and quality varies when the selectivity is the independent variable.

Before "selectivity" can be discussed, it is necessary to know what is to be "selected," and this involves an analysis of the method by which broadcasting is accomplished. The concept of side bands is well known. Broadcasting consists in the simultaneous radiation of two or more high frequency waves. Modulation is the operation of producing high fre-

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quency radiations at a frequency different from a carrier frequency by an amount equal to the audio frequency to be transmitted. The usual methods of modulation produce two such high frequencies, one on either side of the carrier. Thus if a million cycle carrier is modulated by an

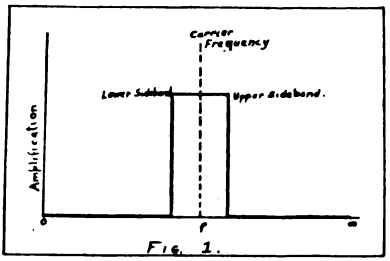

Figure 1

audio frequency of 1000 cycles, three frequencies are radiated, one million minus a thousand, one million, and one million plus a thousand cycles. The frequencies generated by modulation are termed side bands. It should be clearly realized that these side bands are *not* mathematical fictions, but are *physically existent* as a part of the broadcast transmission. It is the "beating" of the side bands on the

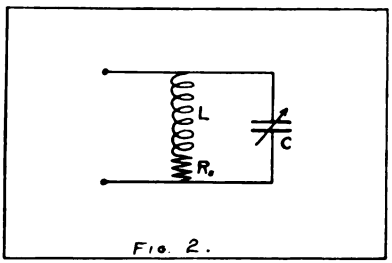

Figure 2

carrier which rectified produces the audio component of the detector output. The two side bands "beating" with the carrier wave add to give the original audio or modulating frequency.

It is obvious that the amplitudes and phases of the side bands must maintain their original relationship throughout any amplification or modifying stages to which the frequency band is subjected.

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Considering the faithfulness of reproduction, or quality, a uniform response curve for all frequencies would be desired. Considering selectivity, this response band should cover only the frequencies included in the side bands of the desired station. Such a response curve is shown in

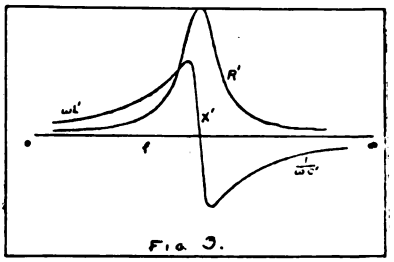

Figure 3

Fig. 1. Below the lower side band and above the upper side band no response is obtained while between these two limits the response is uniform. Such a response curve is obviously a discontinuous function of frequency, and as such requires an infinite series of terms to represent it accurately. This would probably mean an infinite series of selective circuits or circuit arrangements. More will be said later regarding the possibilities of such a system.

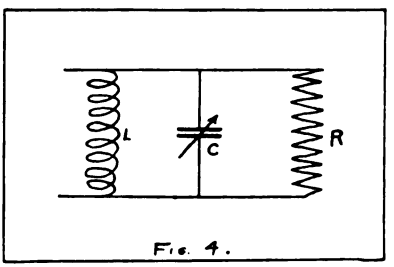

Figure 4

With this conception of the "ideal" response curve in mind, the response curves of the present type of selecting circuits will be more appreciated. The circuit is shown in Fig. 2. The fundamental operation of such a circuit depends upon its change of impedance with change in applied frequency. Below the resonant frequency the circuit is an inductive reactance, and above resonant frequency it is capa-

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city reactance. As resonance is approached, the effective resistance increases, reaching a maximum close to the point of zero reactance. These facts are usually plotted in the form of the curves of Fig. 3.

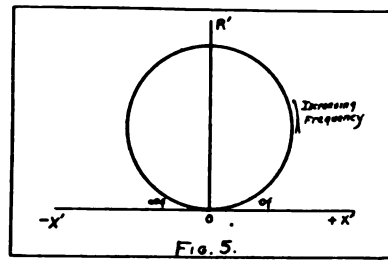

Figure 5

For the purposes of this paper, another viewpoint is preferable. It is a fundamental property of any circuit having a well defined resonant frequency that the curve between effective resistance and effective reactance should be approximately a circle. Thus take the simple circuit of Fig.

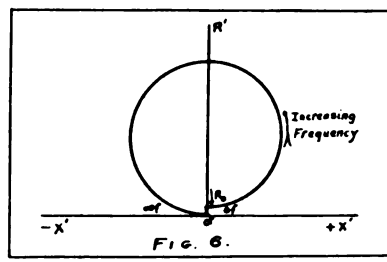

Figure 6

4. The curve between R' and X' is shown in Fig. 5. With zero frequency the inductance "shorts" the circuit. At resonance $\left[f = \frac{1}{2\pi\sqrt{LC}}\right]$ the impedance of the L-C combination is infinite and the diameter of the circle is given by R. At infinite frequency the capacity "shorts" the circuit and thus completes a perfect circle. Changing L and C changes the resonant frequency; changing R changes the diameter of the circle.

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In case there is no shunt resistance, but a series resistance in one, or both legs, the action is somewhat different. Fig. 6 shows the case with a resistance in the inductive leg only. The impedance "circle" begins with a small value R_s, equal to the coil resistance and ends with zero impedance as

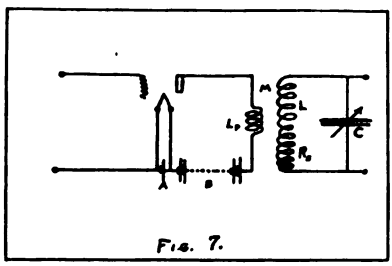

Figure 7

before. The "circle" is not a true circle, but if R_s is small compared with R' it may be treated as such without serious error. In all the cases discussed in this paper this has been done. The case where the error due to this assumption would be greatest has been checked and the error found to be considerably less than 1%, and is sufficiently accurate for all other cases of value to a radio engineer.

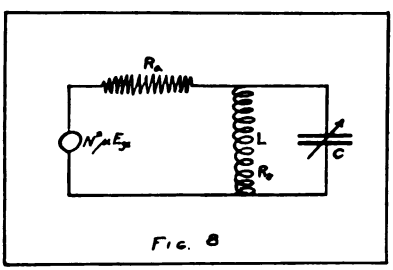

Figure 8

The next consideration is to determine the voltage across such a resonance circuit as the frequency is varied. The usual circuit connections for a vacuum tube amplifier are as shown in Fig. 7. It can be shown that a vacuum tube can be replaced by a generator of zero internal impedance having a voltage μE_{g_1} in series with a resistance R_p equal to the plate-filament impedance of the tube. Also all of the circuit

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constants may be transferred to the secondary by multiplying by N^2. N is the effective transformer ratio, and with no leakage is $\frac{L_2}{M}$. If leakage is present, the transfer is still justified but N may be more difficult to determine numerically.

These changes have been made in Fig. 8, where $N^2\mu E_{g1}$ is the voltage of the transferred generator, and $R_a=N^2R_p$. The vector diagram of the combination impedance is shown in Fig. 9. R_a adds in the direction of R'. The vector AC

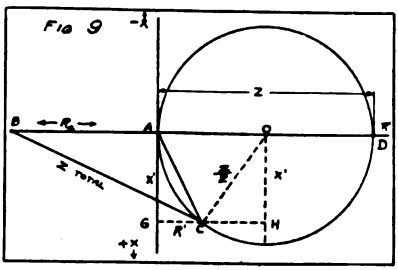

Figure 9

represents the impedance of the tuned circuit and the vector BC represents the impedance of the combination. As the applied voltage divides in direct ratio to the impedances, the response curve is given by the ratio AC/BC as the point C swings around the circle.

$$AC=\sqrt{R'^2+X'^2} \tag{1}$$

$$BC=\sqrt{(R'+R_a)^2+X'^2} \tag{2}$$

Let the response curve ratio $\frac{AC}{BC}$ be represented by B. Then

$$B^2=\frac{R'^2+X'^2}{R'^2+2R'R_a+R_a^2+X'^2} \tag{3}$$

From the geometry of the circle it can be shown that

$$X'^2=ZR'-R'^2 \tag{4}$$

where Z is the maximum impedance of the tuned circuit, ie, the circle diameter, making the substitution of (3) in (4) gives

$$B^2=\frac{ZR'}{R_a^2+R'(2R_a+Z)} \tag{5}$$

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The mathematical solution of the circuit constants (or variables) of a tuned circuit such as shown in Fig. 4 where $R=\frac{L}{R_s C}$ (resonance impedance) gives as the effective resistance a term so that

$$R'=\frac{\frac{L}{R_s C}\ \omega^2 L^2}{\frac{L^2}{R_s^2 C^2}\ (1-\omega^2 LC)^2+\omega^2 L^2} \tag{6}$$

where L, C and R_s are the inductance, capacity and the series resistance of the tuned circuit. At resonance $(1-\omega^2 LC)$ is zero and, as $R'=Z$ at this point,

$$R'_{max}=Z=\frac{L}{R_s C} \tag{7}$$

Substituting (6) and (7) in (5) gives

$$B=\sqrt{\frac{\omega^2 L^2}{(R_a R_s \omega C+\omega L)^2+R_a^2(1-\omega^2 LC)^2}} \tag{8}$$

At resonance $\frac{1}{\omega C}=\omega L$ and (8) simplifies into

$$B=\frac{1}{\frac{R_a}{Z}+1} \tag{9}$$

This means that if the transferred tube impedance is equal to that of the tuned circuit, ie $R_a=Z$, half of the tube generator voltage is effective across the tuned circuit.

This development can now be applied to the determination of selectivity. In this paper the term "selectivity" is used to represent the ratio between the wanted signal and an interfering signal. The usual method of measuring selectivity is based on the width of the resonance curve at half amplitude. While this is a convenient way of comparing two resonance curves, it is meaningless to an engineer who must interpret his solution to fit actual cases. We are interested in the comparative response of a desired station and one that is undesired. Such information is given only by the determination of the response ratios and not by any curve shape. Thus when a circuit is said to have a selectivity factor of five, it means that a wanted signal is five times the strength of an unwanted station, assuming equal inputs. This method has the disadvantage of having a different

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selectivity for every frequency off resonance. However, if all stations are working on a fixed frequency difference basis, such as the *10-kc.* separation of present broadcasting stations, and the selectivity factor is calculated using this difference, the result will be a positive answer to the question of selectivity and not a mere geometrical ratio.

Equation (9) gives the response at resonance, and equation (8) gives the response at any other frequency. ($\omega_2=2\Pi f_2$). Dividing (9) by 8 will give the selectivity of the combination as defined above. Performing this operation and simplifying gives

$$S^2=1+\left(\frac{R_aZ}{R_a+Z}\right)^2\left(\frac{1}{\omega_2L}-\omega_2C\right)^2 \tag{10}$$

At resonance the product terms in (10) disappear and $S=1$. This is obvious from the definition of S. The last term is dependent on the frequency difference from resonance. For a given frequency difference, the only variable is that of $\frac{R_aZ}{R_a+Z}$

This term continuously increases as R_a increases. The value of R_a is the value of the tube impedance R_p transferred to the secondary. *Decreasing* the number of primary turns *increases* R_a and so increases the selectivity. It can be shown that when the primary turns are adjusted for maximum amplification $R_a=Z$. Under these conditions, the selectivity

$$S\propto\frac{Z}{2} \tag{11}$$

If the primary turns be decreased to zero, R_a approaches an infinite value. Substituting $R_a=\infty$ *in* (10) gives

$$S\propto\frac{\infty Z}{\infty+Z}=Z \tag{12}$$

It is thus apparent that decreasing the number of primary turns from the optimum amplification value to an infinitesimal value *cannot more than double the selectivity!*

The increase in selectivity by increasing Z is exactly the same as by increasing R_a. It must be noted however, that while $Z=\frac{L}{R_sC}$ it is not permissable to increase Z by changing the L-C ratio, as this also affects the second product term

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of (10). This effect will be discussed later. However decreasing R_s will increase the selectivity, and if the optimum transformer turn ratio be used, the selectivity will increase as the reciprocal of R_s. Notice that the only way to make the selectivity infinite is for R_s to be zero and the transformer turn ratio to be infinite.

In the above, notice that the square is dropped from S and the product terms. This can be done without error if the product terms are great with respect to 1. As the products terms approach zero, (due to $\frac{1}{\omega_2 L} - \omega_2 C$) the value of S does not differ greatly from 1.

The next problem is to determine how the selectivity changes with the L-C ratio. In order to do this, $\frac{L}{R_s C}$ must be substituted for Z in (10). Simplifying the resulting equation gives

$$S^2 = 1 + \left[\frac{(\omega_r - \omega_2)^2}{\omega_2} \quad \frac{R_a L}{R_a R_s + \omega_r^2 L^2} \right]^2 \tag{13}$$

In considering the action at any large frequency difference from resonance, the bracket term of (13) becomes large with respect to 1. Therefore

$$S = \frac{(\omega_r - \omega_2)^2}{\omega_2} \quad \frac{R_a L}{R_a R_s + \omega_r^2 L^2} \tag{14}$$

The first of these product terms is dependent on frequency off resonance and is independent of circuit conditions. The second term is one which we can adjust. A simple differentiation shows that S is maximum when

$$\omega_r^2 L^2 = R_a R_s \tag{15}$$

But R_a may also be determined by L, for $R_a = \left(\frac{L}{M}\right)^2 R_p$. Substituting this into (15) gives

$$M^2 \omega^2 = R_p R_s \tag{16}$$

This is the value to which M must be adjusted to give maximum amplification. This means that if the secondary inductance is adjusted for optimum amplification, the circuit is also adjusted for *optimum selectivity*, with respect to L-C ratio. This surprising statement indicates that if the circuit conditions are adjusted for maximum amplification, the selectivity is dependent on the L-R_s ratio only. It also means

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that if the circuit is adjusted for maximum amplification of a wanted signal, it will give the greatest ratio between the wanted signal and any interfering frequency.

While this gives optimum selectivity with respect to *L-C*, this is not the maximum possible selectivity. In general the problem is to obtain the greatest (or the least) selectivity for a given amplification. To this end the simplified equation for the amplification of a single tuned stage will be introduced.

$$A=\frac{M\omega \cdot \mu \cdot \omega L}{R_p R_s+M^2\omega^2} \tag{17}$$

Calling $\frac{(\omega_r-\omega_2)^2}{\omega_2}=K_1$, and substituting $R_s=\left(\frac{L}{M}\right)^2 R_p$ in (13) gives

$$S^2=1+\left[K_1\frac{R_p L}{R_p R_s+M^2\omega^2}\right]^2 \tag{18}$$

Solving (17) for $M\omega$, and substituting $A=K_2\mu\omega$ in (18) and simplifying gives

$$S^2=1+\left[K_1K_2^2\frac{2R_p}{L\pm\sqrt{L^2-4K_2^2R_pR_s}}\right]^2 \tag{19}$$

For large frequency differences the *1* may be neglected and with a chosen amplification determining K_2,

$$S\propto\frac{R_p}{L\pm\sqrt{L^2-4K_2^2R_pR_s}} \tag{20}$$

The selectivity will be maximum when the denominator is minimum. If the second term under the radical be zero, the other denominator terms will cancel and the selectivity will be infinite. Notice that this means that R_s is zero, checking a previous conclusion.

The larger L becomes, the more nearly the radical term approaches the value of L, and the smaller the denominator. Maintaining the transformer turn ratio, etc., so as to give a constant chosen amplification, *increasing the secondary inductance will increase the selectivity.* This is a point not so well appreciated by the radio set manufacturer as it should be.

One other point is of interest. For any chosen amplification, i. e. K_2, there are two values of selectivity. One is where

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the amplification is below the optimum value due to too small a mutual inductance, and the other is with the mutual greater than the optimum value. If $L^2=4K_2^2R_sR_p$, there is only one value for selectivity, that which occurs at the optimum amplification.

Before considering the quality factors of this discussion several curves illustrating the previously made points will

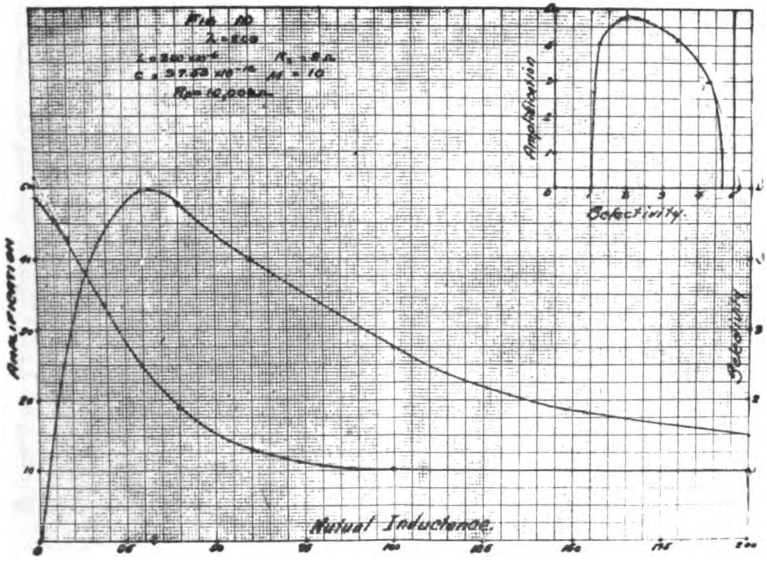

Figure 10

be given. The separate equations (17) and (18) were used in deriving the majority of these curves. In determining the selectivity, K_1 of (18) was calculated on the assumption that the interfering station was 10 kc. off resonance.

The first three curves, Figs. 10, 11 and 12, show the amplification and selectivity curves of a variable condenser tuned stage of radio frequency amplification. The inductance and resistance values chosen are typical of a very good transformer. The average receiver uses less inductance and often has several times 8 ohms in the circuit. These values were chosen to show how poor the selectivity factor is even with such an optimum practical case. Mutual inductance is used as the independent variable, it being customary to vary the primary turns to obtain a desired amplification or selectivity. In the corner of the curve sheets is an insert curve of

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amplification against selectivity. The important point to make here is that with the chosen coil and secondary resistance, every circuit condition *must lie on this curve!* No trick arrangement of capacity or inductance coupling or combination can serve to change the relationship between selectivity and amplification.

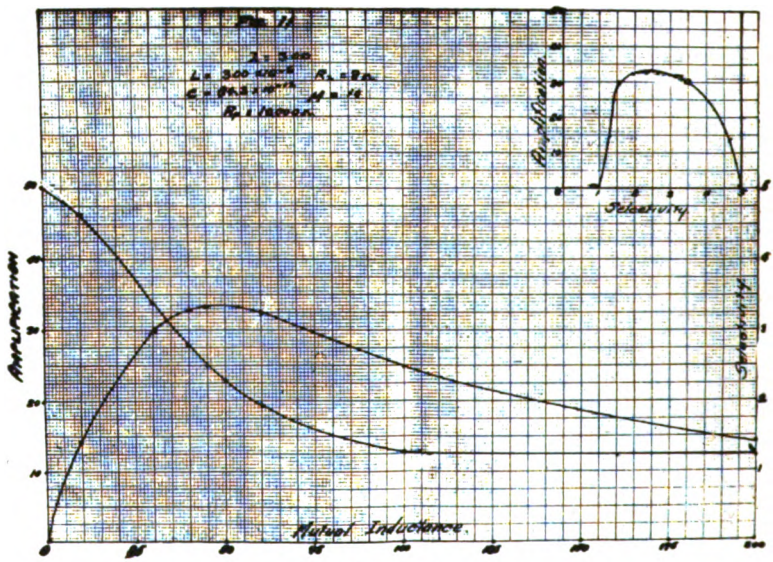

Figure 11

As selectivity is necessary in this type of receiver, a mutual below the optimum value is necessary. Thus in Fig. 10, a mutual of 10 microhenries might be chosen giving an amplification of 30 and selectivity of 4.2. With this fixed mutual the amplification and selectivity are as follows:

	Amplification	Selectivity
200 M	30.	4.2
300 M	14.5	4.6
600 M	4.0	4.8

Take another case. Assume that it is necessary to maintain a constant selectivity. Then we might have

	Amplification	Selectivity
200 M	42.0	3.5
300 M	29.0	3.5
600 M	15.5	3.5

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Or again, assume that some coupling method is used whereby the amplification is constant. The maximum amplification of the 600-meter signal must be the limiting value. Taking approximately this as the constant amplification gives

	Amplification	Selectivity
200 M .	.....16.0	4.6
300 M	.....16.0	4.5
600 M	.....16.0	3.1

The choice of the three methods depends on the class of the set and the difficulties involved in applying the method.

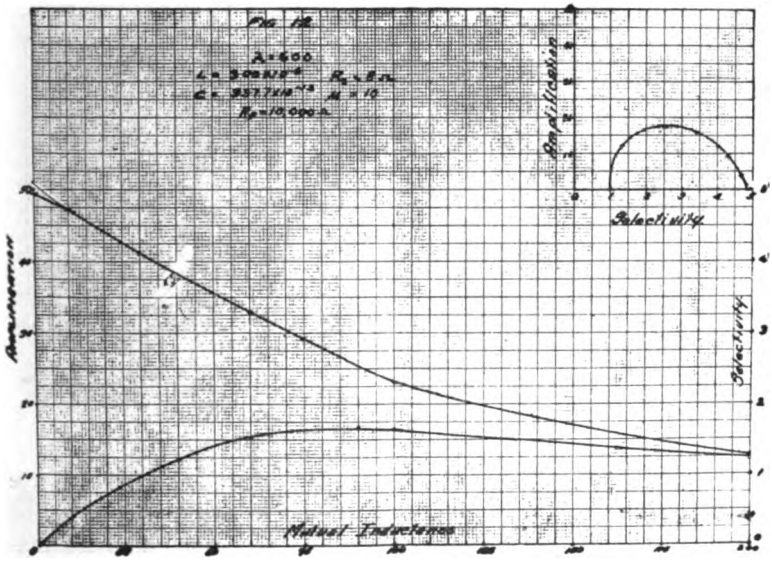

Figure 12

The first method, i. e., constant mutual, is the one which most set manufacturers have been using as it is the simplest. Recently several methods approaching the case of constant amplification have been placed on the market. A compromise between increased amplification and decreased selectivity seems to be the logical outcome of such developments.

Another interesting series of curves are given in Figs. 14, 15 and 16. Here the tuning capacity is left fixed and the secondary inductance varied to resonate the circuit. As it is harder to build a low resistance variometer than a single coil, the secondary resistance was chosen as 20 ohms. Pri-

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mary inductance is chosen as the independent variable in this case, as the changing mutual with change in secondary inductance would not illustrate the usual case clearly.

With this method of tuning both the amplification and selectivity increases with increasing wavelength. A combination of inductance capacity tuning might prove ideal so

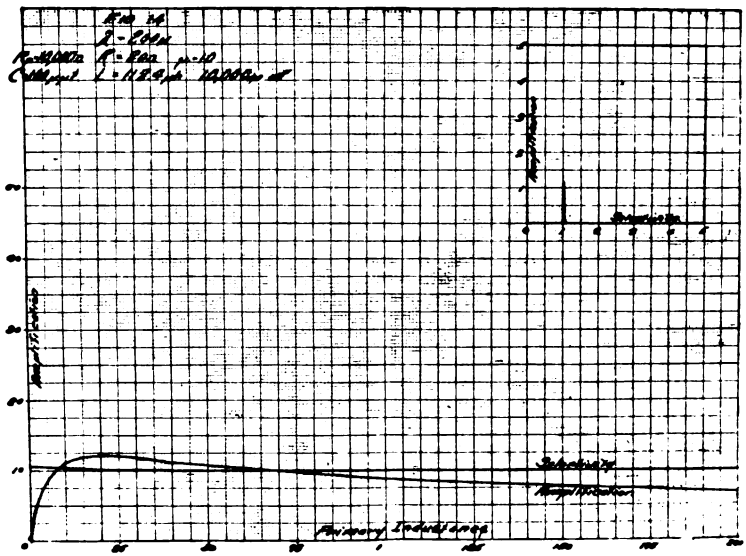

Figure 14

far as present tuning methods are concerned. No tables are offered from these curves as the curves themselves can be read with sufficient accuracy to give any desired information.

Data for amplification and selectivity have been obtained for many other cases and checked experimentally, but the essential facts are given above. Anyone interested in a specific case can substitute his values in Equations (17) and (18) and quickly determine the curves.

The discussion thus far has considered only the relation between selectivity and amplification. As was pointed out in the introduction, the quality of reproduction is affected by these other variables and must also be considered. There are several factors which affect the quality, each by a different amount. The major factor is that produced by the non-uniform amplification of the desired side bands.

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In this discussion, percentage quality is defined as the area below the response curve and between the side band limits, divided by the area of the rectangle whose length is given by the maximum amplification and whose width is the side band separation. This meaning of quality assumes that all frequencies between the two side bands are present, and

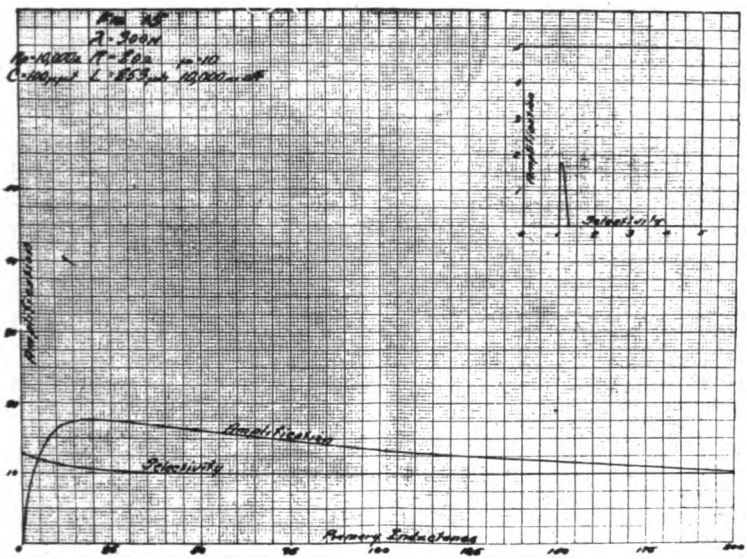

Figure 15

that the energy at each frequency is of equal importance. This may be too severe a restriction as undoubtedly some of the higher audio harmonics can be dropped without materially affecting the quality as perceived by ear. There seems to be no logical basis for making any other method of quality measurement involving a frequency spectrum, and it is to be hoped that some such method can be universally adopted. It might be preferable to plot both the amplitude and the side band frequencies (on both sides of resonance as a base line) in logarithmic units and measure the area in the customary square units. In any case, a fair definition of quality seems to be the area of the received frequency spectrum divided by the area of the ideal frequency spectrum having the same maximum amplification.

In a tuned radio frequency amplifier the grid to grid voltage amplification is increased in two places. First in the

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station's interference. The ratio of the area of the desired station's signal (actual case, not ideal) to that of the interfering station is a measure of the selectivity of the circuit. In the tables following this selectivity factor is called S_A, as it is based on the respective area. The selectivity factor as defined in the first part of this paper, i. e., the ratio of the resonant frequency signal to that 10,000 cycles off reso-

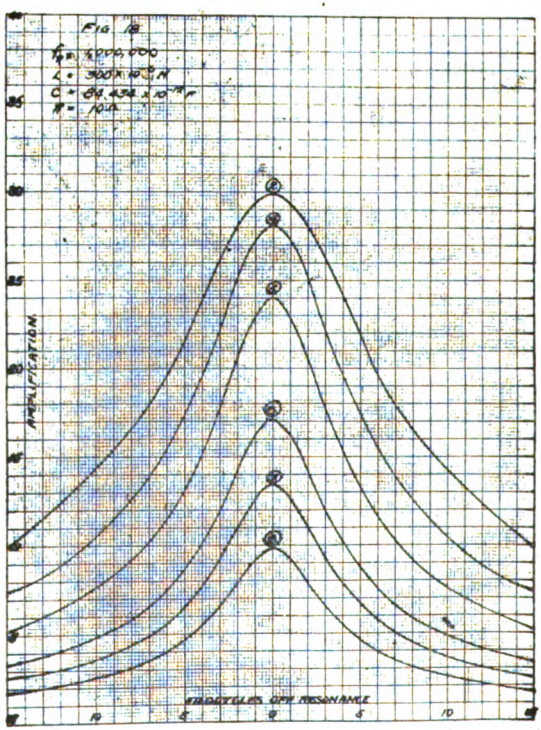

Figure 18

nance is called S_o as it is based on respective ordinates. It is obvious that S_o will be larger than S_A. S_A is approximately S_oQ where Q is the per cent. quality, or quality factor as previously defined.

As the circuit is made sharper, both the interference and the quality decrease. The ratio between quality and interference is also given in the tables. Notice that for a given resistance circuit the ratio does change much with M, although increasing both with decreasing R and increasing number of stages.

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The second series of curves are those with a constant secondary resistance and varying mutual. Some of the values of mutual were chosen simply to make the amplification of the product stages equal.

Several curves were made of product stages and the data given were calculated from the shape and area of these curves. The curves themselves are not given as they would add nothing to the discussion.

TABLE 1. (Fig. 17)

Curve No.	Max. Amp.	M.	R_s	Q%	I%	Q/I	S_A	S_o
1	42.2	35.7	5	73.1	29.6	2.47	2.7	3.8
2	29.8	50.8	10	89.2	54.8	1.63	1.8	2.1
3	18.8	79.6	25	98.0	82.0	1.20	1.2	1.2

TABLE 2. (Fig. 18)

Curve No.	Max. Amp.	M.	R_s	Q%	I%	Q/I	S_A	S_o
2	29.8	50.8	10	89.2	54.8	1.63	1.8	2.1
4	28.1	35.8	10	84.0	46.9	1.83	2.1	2.7
5	23.8	25.4	10	79.4	42.9	1.85	2.3	3.1
6	17.1	16.1	10	75.9	40.6	1.87	2.5	3.5
7	13.5	12.1	10	75.0	39.5	1.90	2.5	3.6
8	9.8	8.6	10	74.8	38.5	1.94	2.6	3.9

TABLE 3

Curve No.	Obtained By	Max Amp.	Q%	I%	Q/I	S_A	S_o
9	6 x 6	293.	63.0	16.7	3.77	6.0	12.2
10	4 x 8	293.	67.0	21.9	3.09	4.6	7.8
11	5 x 5	567.	66.0	18.5	3.17	5.4	9.6
12	1 x 7	568.	58.8	14.8	3.97	6.7	13.5
13	5 x 5 x 5	13,114.	55.6	8.7	6.38	11.6	28.8
14	5 x 5 x 5 +2000, 0 and 2000, off resonance	10,000	64.0	10.6	6.04	9.4	22.0

The curve 14 of Table 3 is very interesting. This was made with the first stage tuned to 2000 cycles above the signal frequency of 1,000,000, the second stage was resonant and the third stage was tuned 2000 cycles below resonance. This corresponds to a capacity variation of approximately 4%, about that value to which most good variable condensers are held. The surprising thing is the fact that this curve is so little different from that of 13, where all stages are resonant. Apparently such a variation does not seriously affect the general characteristics.

It would be unwise to leave the subject of selectivity without some mention of regeneration, for regeneration has long been considered one of the most satisfactory means of

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station's interference. The ratio of the area of the desired station's signal (actual case, not ideal) to that of the interfering station is a measure of the selectivity of the circuit. In the tables following this selectivity factor is called S_A, as it is based on the respective area. The selectivity factor as defined in the first part of this paper, i. e., the ratio of the resonant frequency signal to that 10,000 cycles off reso-

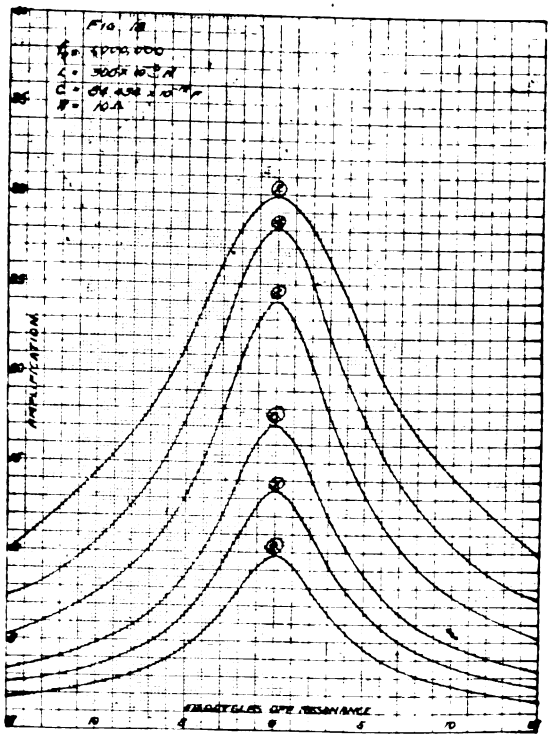

Figure 18

nance is called S as it is based on respective ordinates. It is obvious that S will be larger than S_A. S_A is approximately SQ where Q is the per cent. quality, or quality factor as previously defined.

As the circuit is made sharper, both the interference and the quality decrease. The ratio between quality and interference is also given in the tables. Notice that for a given resistance circuit the ratio does change much with M, although increasing both with decreasing R and increasing number of stages.

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The second series of curves are those with a constant secondary resistance and varying mutual. Some of the values of mutual were chosen simply to make the amplification of the product stages equal.

Several curves were made of product stages and the data given were calculated from the shape and area of these curves. The curves themselves are not given as they would add nothing to the discussion.

TABLE 1. (Fig. 17)

Curve No.	Max. Amp.	M.	R_s	Q%	I%	Q/I	S_A	S_o
1	12.2	35.7	5	73.1	29.6	2.47	2.7	3.8
2	29.8	50.8	10	89.2	54.8	1.63	1.8	2.1
3	18.8	79.6	25	98.0	82.0	1.20	1.2	1.2

TABLE 2. (Fig. 18)

Curve No.	Max. Amp.	M.	R_s	Q%	I%	Q/I	S_A	S_o
2	29.8	50.8	10	89.2	54.8	1.63	1.8	2.1
4	28.1	35.8	10	84.0	46.9	1.83	2.1	2.7
5	23.8	25.4	10	79.4	42.9	1.85	2.3	3.1
6	17.1	16.1	10	75.9	40.6	1.87	2.5	3.5
7	13.5	12.1	10	75.0	39.5	1.90	2.5	3.6
8	9.8	8.6	10	74.8	38.5	1.94	2.6	3.9

TABLE 3

Curve No.	Obtained By	Max Amp.	Q%	I%	Q/I	S_A	S_o
9	6 x 6	298.	63.0	16.7	3.77	6.0	12.2
10	4 x 8	293.	67.0	24.9	3.09	4.6	7.7
11	5 x 5	567.	66.0	18.5	3.17	5.1	9.6
12	1 x 7	508.	58.8	14.8	3.97	6.5	13.5
13	5 x 5 x 5	13,144.	55.6	8.7	6.38	11.6	28.8
14	5 x 5 x 5 +2000, 0 and −2000, off resonance	10,000	61.0	10.6	5.01	9.4	22.0

The curve 14 of Table 3 is very interesting. This was made with the first stage tuned to 2000 cycles above the signal frequency of 1,000,000, the second stage was resonant and the third stage was tuned 2000 cycles below resonance. This corresponds to a capacity variation of approximately 4%, about that value to which most good variable condensers are held. The surprising thing is the fact that this curve is so little different from that of 13, where all stages are resonant. Apparently such a variation does not seriously affect the general characteristics.

It would be unwise to leave the subject of selectivity without some mention of regeneration, for regeneration has long been considered one of the most satisfactory means of

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increasing the selectivity of a tuned radio frequency amplifying circuit. If it is assumed that all of the vacuum tube parameters are constant or linear functions of the grid voltage, it may be shown that the power fed back by the tickler is equal to that which would be drawn from the amplifier tube ahead if the tuned circuit resistance were

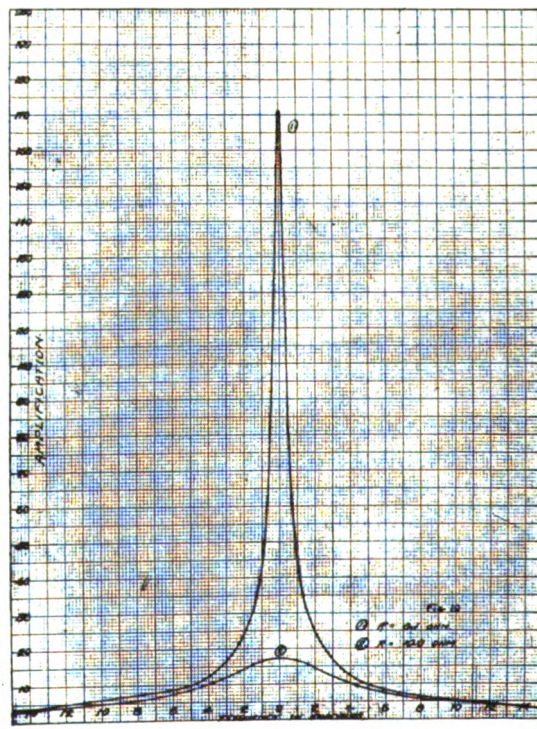

Figure 19

lowered. This fact has lead to the conception of regeneration as "negative resistance". It must be emphasized that the only basis for considering regeneration as the introduction of negative resistance into the tuned circuit, lies in the truth of the above assumption. This assumption is far from true when the "negative resistance" approaches the positive resistance in value. However, at values of "effective resistance" large with respect to infinitestimal values, certain calculations may be made without appreciable error.

A resonance curve for the case of Curve No. 6 where

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the effective resistance has been reduced to *0.1* of an ohm is plotted in Fig. 19. Notice that the low resistance of the secondary (constant mutual in both cases) does not produce *decreased* interference, but only an *increase* in signal amplitude, particularly in the range close to resonance. This of course increases the selectivity factor as previously defined, but also hurts the quality. These facts may or may not help the receiver, depending on other conditions.

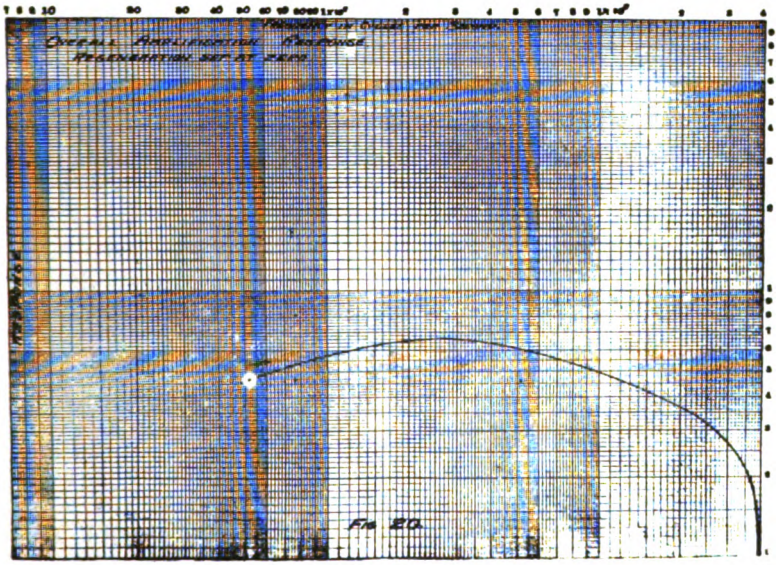

Figure 20

Two other points should be considered as affecting the quality. The first is the phase shift of the side bands as they pass through the amplifier. The peak of the "beat" between the lower side band and the carrier will not coincide with the "beat" between the upper side band and the carrier. If the receiver is tuned to resonance, these audio beats are symmetrical with the resonant period and add to give a single peak which is identical in phase but decreased slightly in amplitude from the original. If the receiver is not tuned to resonance the audio components will be shifted slightly with respect to each other, but as shown by Fletcher and others, a phase shift of as much as a cycle does not seem to affect the ear. The phase shift can therefore apparently be neglected as affecting quality.

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The second factor is the transient response of the circuit. The previous equations dealt only with a constant amplitude, continuously applied frequency. In broadcasting, the amplitude variation of each single frequency must be followed. If the decrement of the receiver is too low, the grid voltage will not be able to follow the amplitude variations of the signal. This problem has been considered in detail elsewhere and will not be stressed here. A typical case where $L=300\times10^{-6}$ *hy*, and $R_s=10$ *r* was calculated and (neglecting additional damping of coupled plate resistance) it was found that only *0.00014* seconds were required to reduce the grid voltage to 1% of its original value. The additional damping of the coupled plate resistance brought this time down to about *0.00010* seconds, corresponding to a variation frequency of 10,000 cycles. As an alternating frequency current cannot well change in amplitude faster than its own frequency, and the lower limit shown above is approximately 10,000 cycles, the quality of present day receivers will not be much affected by this factor.

A consideration of the three factors of amplification, selectivity and quality leads directly to an old problem of design. Shall each step or unit in the instrument be as near perfect as possible to make it, or shall one unit attempt to compensate for the deficiency of another? Many arguments can be advanced both ways. Certainly the ideal way would be to have each unit perfect in itself. Is this possible commercially?

Fig. 20 shows the most interesting curve in this whole paper. The curve is the response curve for a well known 5-tube receiver, incorporating two stages of radio frequency amplification, a regenerative detector and two stages of audio frequency. The tuned circuits were designed, principally by adjusting the secondary resistance and transformer mutual inductance, to be quite selective and still maintain good amplification. The audio transformers were somewhat better than the average, although beginning to fall at 200 cycles and falling very rapidly below 90 cycles. Above 200 cycles, the audio transformer curves were practically flat to 6000 cycles and then rose slightly to 9000 cycles where they cut off sharp.

The regeneration was not critically adjusted, yet served to keep the amplification at low frequency (at least to 40 cycles) very good. Here was one case where peaked radio

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frequency amplification compensated for poor audio amplification. However, the sharpness of the radio frequency stages began hurting the audio response curve at about 500 cycles. At 2000 cycles the response falls very fast and at 4000 cycles no signal was obtained. The beautiful "flat" transformer characteristic above 500 cycles was absolutely useless. Considering 5000 cycles as the side band limit, the audio transformers should have had a peak at about 4000 cycles.

This discussion shows that even the audio transformers affect materially the selectivity of the receiver, and based on present ideas in set construction and present type selecting circuits, *should never have a flat top characteristic.* This point cannot be too strongly emphasized, for if we do not make *each* unit of our receiver almost perfect, there is no logic about only making one unit perfect. It is better to make one unit compensate for another, including everything from the antenna to the reproducing unit. This has always been the commercial way of getting results desired in the quickest time, and considering the engineering handicaps is often the best way.

However, in the long run, the ideal way will prove the best; that of making each unit perfect. How shall this be done? If the answer were known it would be done now. One statement is clearly true, however. In the ideal receiver, selectivity, amplification and quality will be absolutely independent. Selectivity may come through properly designed band pass filters. Amplification will be easily controlled by the operator and independent of frequency. With true aperiodic amplification, no phase distortion and high decrement pickup circuits, quality will be perfect. The road ahead of the radio engineer striving to build the ideal receiving set is long and hard.

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RADIO PHENOMENA RECORDED BY THE UNIVERSITY OF MICHIGAN GREENLAND EXPEDITION—1926

BY

P. C. OSCANYAN, JR.

(In Charge Expedition Radio)

Radio, and especially short wave radio, has found its place in the plans of almost every expedition which has left the United States within the past year, and because such radio equipment is being carried into the more or less inaccessible places of the world we are learning that the explored places are, in many cases, different from what we had expected.

The University of Michigan Greenland Expedition of 1926, directed and led by Prof. W. H. Hobbs, was the Expedition of which I was a member and in charge of radio. The Expedition went in search of meteorological data, to lay down a base station for work in 1927, and, in connection with the latter work, to determine a proper location for a short wave radio station to communicate with the Schooner Morrissey and the United States.

A base camp site was selected some 50 miles East of Holstensborg, on Maligiak Fiord, North of the Arctic Circle, Reference to the photograph will show that the camp was situated in the center of a natural bowl-like valley and it was close to the location of the tents that the radio station was first set up. Take careful note of the figures which are written in above the various high points which surround the camp.

At this point it is necessary that we digress for a moment. Last spring John L. Reinartz, in a speech before the A. R. R. L. Convention in New York City, brought to our attention, for the first time, the fact that when a radio receiving station which plans to work on wavelengths of 50 meters or below is placed at the foot of a hill or mountain which is of a height greater than 17 degrees from the hori-

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zontal of the station, then signals will be screened off from the receiver.

It was with a view to testing this that the station was first established at that point. The results were so prompt that they were thoroughly convincing. In order that we might be doubly sure of the height of these high points we

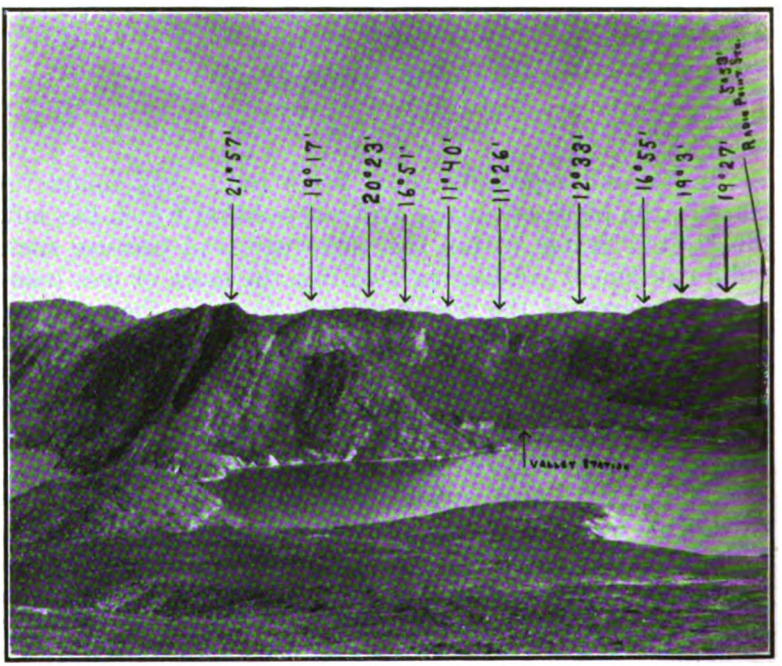

Photo by Prof. L. M. Gould

Figure 1. Figures and arrows indicate heights of hills surrounding the station in the valley.

(Mr. R. L. Belknap, Expedition Surveyor, and I) checked them with a theodolite which was placed midway between the mast and the receiving set, a matter of some 25 feet. The angles given in the illustration are the angles thus obtained.

Finding ourselves screened off from signals coming from the North, thus being cut off from the vessel which had taken us up and was to take us back after making arrangements by radio, and also being cut off from all but the most powerful short wave stations in Eastern and Central United States another location less sheltered was determined upon

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for the radio station. It is interesting to note here that in all probability due to the clearing effect of the screening, high points stations in the western U. S., in an unscreened direction from us, and stations in Europe and even New

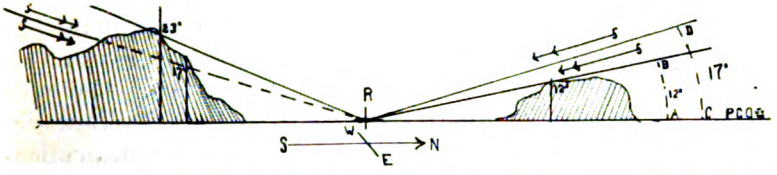

Figure 2. Outline illustration of method of obtaining angles and the blocking of signals by the surrounding territory.

Zealand, also in an unscreened direction, were logged having good strength whereas when our later location was reached these signals were, in all probability, drowned out by signals from stations closer by.

Photo by Prof. L. M. Gould

Figure 3. Cross at extreme left indicates location of Station on Point of land 400 ft. high and projecting out into fiord.

Relocating our station at point B, named Radio Point, gave a prompt and satisfying return to what would be the normal condition expected for that locality. It would also appear that static in that part of the world is the overflow

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from the more harassed lower latitudes, it not being present in sufficient quantity to cause notation in my log until after we had shifted our station to the higher point.

Another interesting point of note is that the distance within which "brute force" signals may be heard is greatly increased at sea. By brute force signals, I mean the signals which may be received from a short wave transmitter within a limited area surrounding the station. The radiated waves which follow the earth or sea surface die out within a comparatively short distance due to absorption.

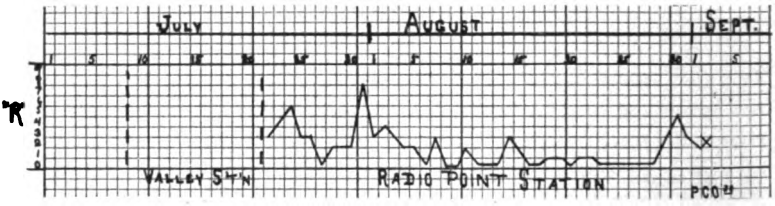

Figure 4

The thing which these experiments bring most forcibly to my mind is that such a critical angle of screening of short wave signals rather sets at odds the theory laid down by Dr. A. H. Taylor that there is a secondary, or even tertiary skip distance. The angle at which these short waves are screened out, points, as Mr. Reinartz has shown me, to the angle of refraction of light on pure water.

Now it is a known fact that dry air is heavier than moist and that as we ascend in the atmosphere there is more and more moisture present so that when the level of the cirrus cloud is reached we find that they are moisture frozen to minute ice particles, (such at least is the general condition). It is also true that this moist band of atmosphere experiences a diurnal change and is higher at the equator than at the poles, all of which would seem to point to the hypothesis that, if this 17 degree angle is at all significant it would warrant the assumption that high frequency radio signals are reflected by this moist and ionized upper atmosphere.

If the foregoing is true, then it would seem that unless the signal struck a reflecting surface of like character, when it reached the earth, it would never be sufficiently reflected so that it might rise again through the atmosphere

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to the Kennelly-Heaviside layer with sufficient strength to permit it to again appear at another point on the surface of this earth.

The popular idea that there is no static in the Arctic will receive a setback when the static audibility graph given herewith is consulted. I would greatly appreciate it if any

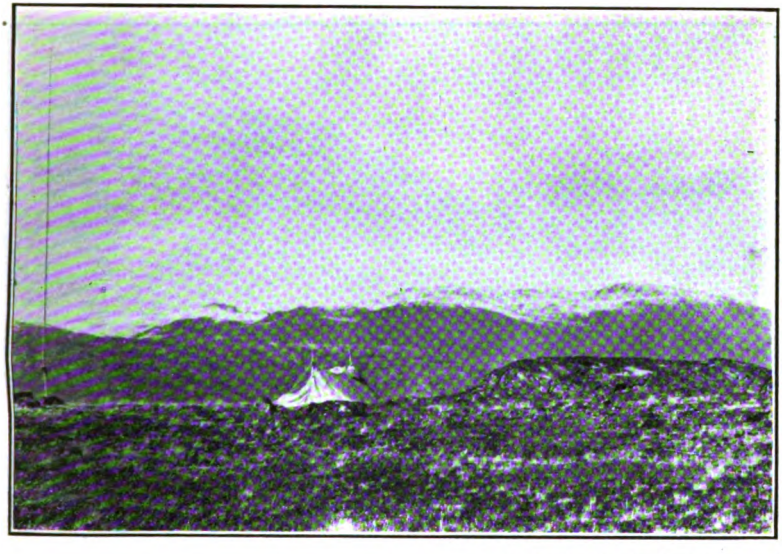

Photo by Prof. L. M. Gould

Figure 5. Radio station on the Point. Shelter tent housing apparatus and operator. Mast is sectional bamboo 35 ft. high. View N. W.

among those who read this article can give me any information as to whether the high audibility static on certain dates either precedes, accompanies, or follows, static of the same or greater intensity elsewhere in the world. A comparison of static recordings daily during the period shown would be of great interest to me.

As mentioned before, there is to be another expedition next year and it is planned to remain up there all during the winter of 1927-28. I appreciate that this will be an excellent opportunity for me to cooperate with anyone who wishes to collect comparative data or would like to suggest some particular field of radio for us to explore, since the radio station is to be a part of the expedition all during that time.

The antenna system consisted of a single wire 40 feet

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long suspended from a sectional bamboo mast 35 feet high. This mast was of 2 inch bamboo and in 5 foot sections, for strength and portability, guyed with stranded antenna wire, Pyrex insulators, and steel tent pegs. The counteerpoise consisted of 3 wires each 20 feet long, No. 14 rubber covered, laid out fanwise on the ground.

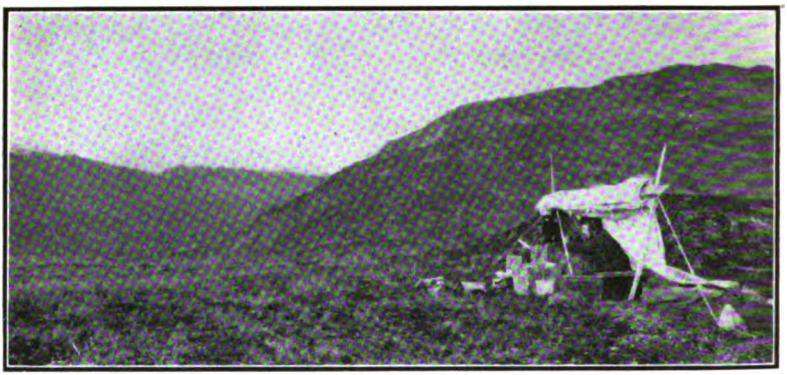

Photo by Prof. L. M. Gould

Figure 6. Looking North over the Station on the Point. Shelter tent rolled back to show method of housing and operation.

The receiver was especially made for the expedition by the Burgess Laboratories at Madison, Wis. The transmitter was a homemade one employing two Cunningham CX301A tubes. Using only 200 volts on the plates of these tubes we were able to communicate over distances greater than 1000 miles. The receiving circuit was a modified Reinartz, the transmitting circuit was a shunt feed Hartley. Detector and one step a. f. were sufficient for reception.

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PUNCTURE DAMAGE THROUGH THE GLASS WALL OF A TRANSMITTING VACUUM TUBE*

By
YUJIRO KUSUNOSE

(Electrotechnical Laboratory, Hunisky of Communications, Tokyo, Japan)

INTRODUCTION

Puncture damage through the glass wall has rarely been met with radio transmitting vacuum tubes which are under operation at ordinary long wavelengths. However since short wave radio has recently come into use, frequent occurrence of this damage has been noted, which leads us to suspect the existence of some special causes other than those previously known (such as insulation breakdown of the glass wall or melting of the glass due to high ohmic losses at high temperature.) The causes have been investigated by the writer and are attributed to the dielectric loss in glass, as described below.

TYPE OF DAMAGE

The phenomenon was first encountered at the Hiraiso Branch Office of the Laboratory, when the Marconi valve MT-4 (anode 10 kv. 200 w.) was being used as a short-wave oscillator. That such a damage is due to dielectric loss may be seen from the conditions of its occurrence, viz., (1) the punctures always appear somewhere on the neck of the glass bulb on the side of grid stem, and marks of softening are observed on the glass wall around that part (Fig. 1), which may be considered to be region of most intense radio frequency electric field, and (2) the breakdown occurs at the instant when operating efficiency is high and anode loss rather low.

THEORETICAL CONSIDERATIONS

In the first place, the matter will be considered from the theoretical point of view and thus it will be found that the

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softening of glass is due to dielectric loss. When a tube is oscillating, the anode and grid voltages are always opposite in sign, and therefore radio frequency alternating potential difference amounting to 10 kilovolts may exist between the two electrodes. Assuming the distribution of electric field built by the potential difference as depicted in Fig. 2 which

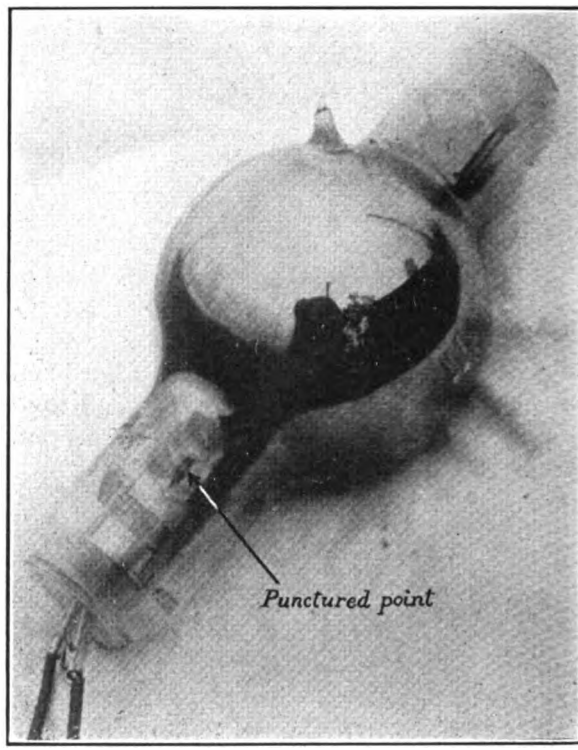

Figure 1

shows the construction of above mentioned valve MT-4, the most intense field may be supposed to exist around the neck of the bulb marked *b*, and is calculated to be of the order of 1 kv./cm.

The dielectric loss in glass as approximately

$$W = k F^2 f \qquad watts/cm^3$$

where F is the field strength in kv./cm. mentioned above, f the frequency in kilocycles, and k the factor depending on the material. The value of k for lead glass as used in the

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valve MT-4 has been found to be $k=19\times10^{-4}$, and assuming $f=15\times10^{3}$ *kc.* ($\lambda=20m.$) the loss at the above-mentioned part of the glass wall may be calculated from the formula to be 0.36 watt/cm³. The thickness of the glass being 0.17

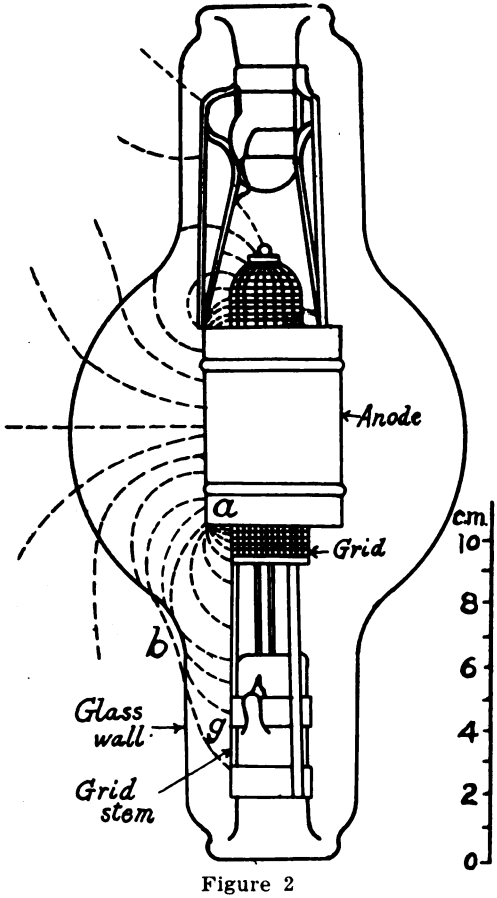

Figure 2

cm., the loss per unit of surface area becomes 0.061 watt/cm².

The temperature rise on this part of glass wall may be roughly estimated in the following manner. Under full load conditions, the valve MT-4 consumes 80 watts in the filament and dissipates 200 watts in the anode, and thus 280 watts in total are radiated as heat through the whole glass surface of 450 cm²; i. e. 0.62 watt/cm² is radiated in

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average. Assuming the fraction of the heat to be absorbed by the glass out of the radiation to be 20%, 0.12 watt/cm² will be lost in the glass. Actual observation under the same load condition has shown that the temperature of the glass wall reached 100 to 160°C.

Now that a dielectric loss of the order of 0.06 watt/cm² is added to the above under conditions of short wave opera-

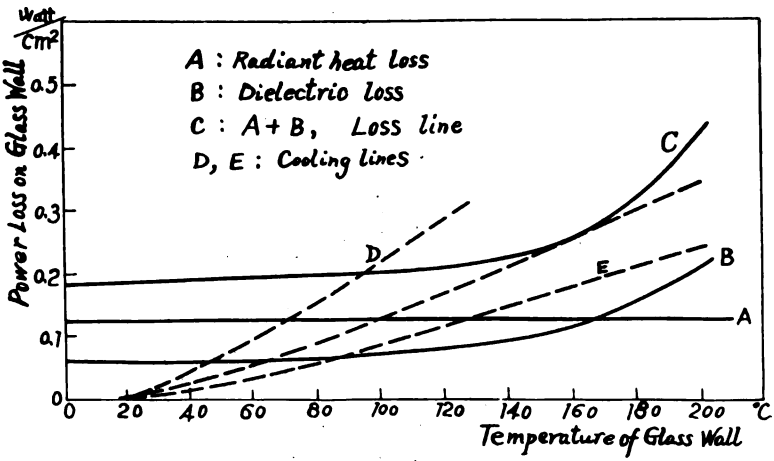

Figure 3

tion at full load, the final temperature may be estimated to reach 200°C or thereabout. This is, however, somewhat lower than the softening point of the glass (about 400°C) and cannot, therefore, explain the damage.

The above reasoning is based on the assumption that dielectric loss is independent of temperature. But this may not be the case; and therefore the variation of the factor k with temperature has been observed and the following results were obtained which show that the dielectric loss increases very rapidly with temperature at the higher temperatures.

Temp.	20	60	100	140	170	200	°C
k	19	20	24	31	43	69	$\times 10^{-6}$

From these measurements the softening of the glass may be explained in the following way. In Fig. 3 the abscissae represent the temperatures of the part of the glass wall under consideration, and the ordinates the amounts of power loss that take part in heating the glass. The heat

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radiated from the electrodes and absorbed in the glass is independent of temperature, and may be represented by a horizontal line A (0.12 watt/cm^2.)

The dielectric loss in the glass is 0.06 watt/cm^2 at room temperature, but increases with temperature in proportion

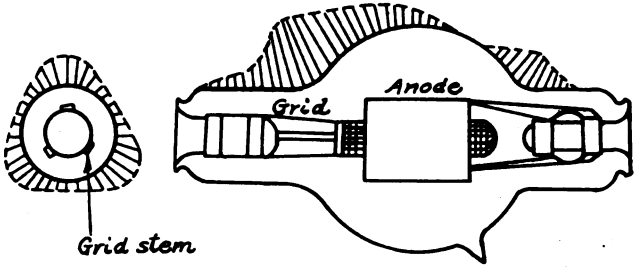

Figure 4

to the value of k in the above table. This may be represented by a curve B. Thus the total amount of power heating the portion of the glass is the sum of A and B which makes the "loss line" C.

On the other hand, the amount of heat which escapes from the glass wall increases with temperature. The dotted lines which are nearly straight represent these "cooling lines". The less effective the cooling, the less will be the inclination of the line.

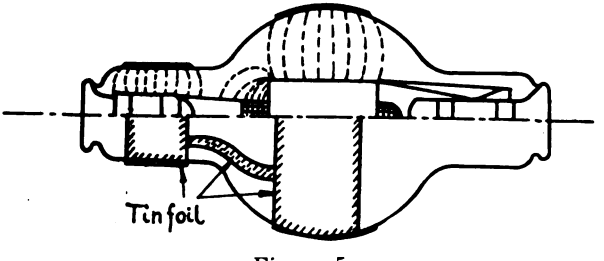

Figure 5

The final temperature reached by the glass is then indicated by an intersecting point of the "loss line" with the "cooling line", as shown D in Fig. 3; but if cooling is so poor that the "cooling line" does not at all meet the "loss line" such as E in the figure, the loss in the glass is by no

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means fully dissipated and the temperature rises indefinitely, and softening or melting of the glass eventually takes place. As a consequence of this cumulative effect and the lack of uniformity in heating and cooling of glass, some part of the glass begins to melt, which results in a puncture because of external air pressure.

EXPERIMENTS

1. The valve MT-4 was suspended horizontally and a radio-frequency voltage of 5,000 V. at 2,500 kilocycles was

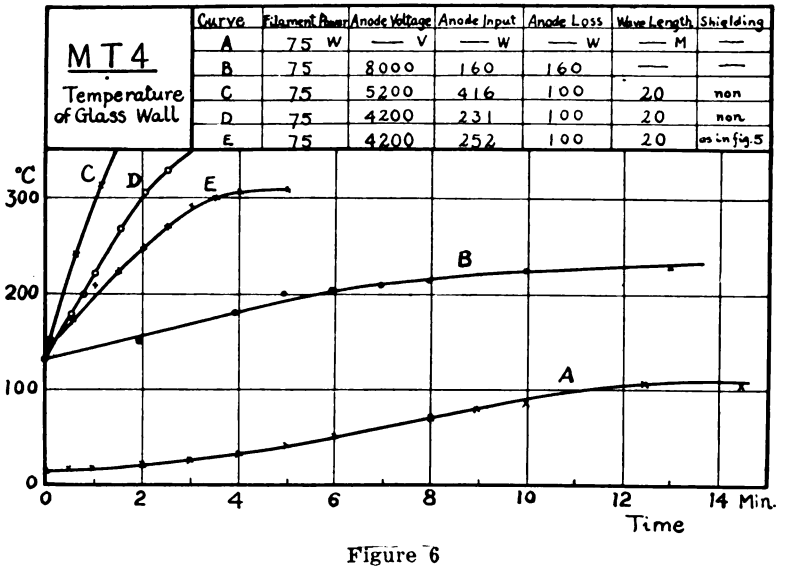
MT4
Temperature of Glass Wall
Curve | Filament Power | Anode Voltage | Anode Input | Anode Loss | Wave Length | Shielding
A | 75 W | — V | — W | — W | — M | —
B | 75 | 8000 | 160 | 160 | — | —
C | 75 | 5200 | 416 | 100 | 20 | non
D | 75 | 4200 | 231 | 100 | 20 | non
E | 75 | 4200 | 252 | 100 | 20 | as in fig.5
°C
300
200
100
0
0
2
4
6
8
10
12
14 Min.
Time
A
B
C
D
E

Figure 6

applied between the anode and grid, the filament not being lighted. The temperature of the glass bulb increased and finally attained the distribution illustrated in Fig. 4, the temperature around the neck of the bulb being the highest and reaching about 100°C. When the bulb was shielded by two sheets of tin foil connected electrically as shown in Fig. 5, no heating of this part could be perceived.

2. The same valve was put into operation as a short wave transmitter, and the temperature at a point on the neck of the bulb was measured. The data are plotted in Fig. 6. When the tube was oscillating at short wavelengths, the temperature rose quite high and sometimes had a tendency

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to increase rapidly, even beyond 300°C, which would indicate the likelihood of damage resulting from continuation of operation.

Prevention of Damage

As the cause of the damage has been found to be the dielectric loss in glass, it is simply remedied; and a knowledge of the causes of puncture will also be helpful in securing high efficiency of operation.

The methods of prevention may be summarized as follows:

A. On operation of the tubes:

1. Thoroughly cool that portion of the glass wall which is likely to be exposed to intense radio-frequency electric fields.

2. Shield the glass wall electrostatically so as to prevent any portion of it from being exposed to an extremely intense electric field.

3. Carefully support the tube so that no part of supports may prevent meeting the above conditions.

4. Lower the anode voltage and output load from the rated values suitable for operation at long wavelengths to values which will keep the temperature of the portion within safe limits.

B. On designing of the tubes:

1. Use low-loss and heat-resisting material for the bulb.

2. Choose such shapes of the bulb and electrodes that strong stray electric field may not reach the walls.

3. Electrostatically shield the electrodes so as to avoid stray electric fields extending to the glass wall.

C. On manufacturing the tubes:

Continue evacuation of the tube after the usual bombardment, operating it meanwhile as a short wave oscillator.

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DISCUSSION ON FIELD DISTRIBUTION AND RADIATION RESISTANCE OF A STRAIGHT VERTICAL UNLOADED ANTENNA RADIATING AT ONE OF ITS HARMONICS* (S. A. LEVIN AND C. J. YOUNG)

O. C. Roos: Two names stand out in the early work on antenna radiation at shorter than fundamental wavelengths: John Stone Stone of Boston, and F. Hack, a German. Since neither of these early investigators was mentioned in the paper of Messrs. Levin and Young in Vol. 14, No. 5, of the Proc. of the I. R. E., the present memorandum is undertaken as a contribution to the history of the early aspects of the subject.

Stone was the first investigator to study the radiations of a simple ungrounded antenna, taking into account the image of this antenna. In doing this he introduced the then comparatively new operational method of handling differential equations, discovered by Oliver Heaviside, from whom he received a grateful letter of thanks.

I had the privilege of reducing the latter part of this work to a form suitable for engineering use and in particular the differential equations for a receiving antenna, first given out at St. Louis in 1904. Their solution in 1907 by myself showed the possibility of the receiver vertical, vibrating at harmonics which were even multiples of 1, 3, 5, etc. As was shown in 1902 by Stone, a transmitter could not vibrate at any even multiple of its fundamental frequency.

The problem of radiation was considered by Stone to be that due to a double sheet of oscillating current elements of different length. One sheet was below the ground and was the image of the other. This was the first treatment of an oscillating Hertz "image" and was made public in the Electrical Review of October 15, 1904 and applied publicly on July 4, 1908 in a "Review" paper on "The Resistance Equivalent of Radiation."

*Received by the Editor November 2, 1926

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In this paper the field at a distance of 200 or more wavelengths, was obtained as is shown in Fig. 1.

In Fig. 1-A we have a Marconi wire a-t as a simple radiator driven at the base by a radio frequency alternator with one lead grounded. The frequency is 3/2 of the fundamental, hence a quarter wave is 2/3 the antenna height or c-t.

If the current at the base is called unity and is proportional to the horizontal a-b, then the maximum current at CD is proportional to cosecant 2/3 of 90 degrees or csc

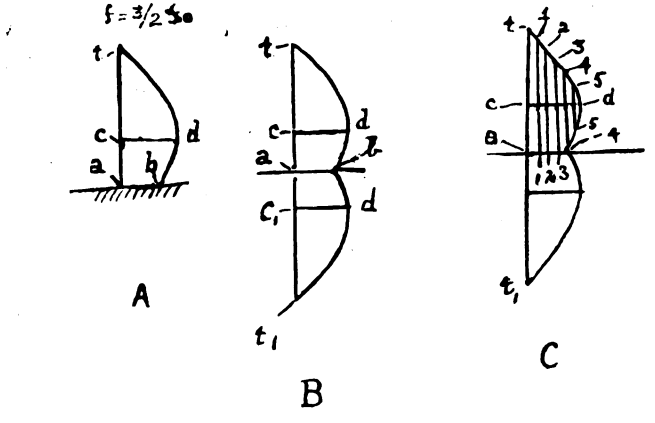

Figure 1

60 degrees. It is 1.16 approximately or 16% greater than the current at the base of the antenna.

In Fig. 1-B both the Marconi wire and its image a-ti are shown with their identical current distributions—both positive or adding their effects in producing radiation. However, at a great distance from "a" these effects are assumed quite practically, to be equal.

The field strength at a distant point is nearly doubled by the image and hence the mathematical determination of this field can be carried on by considering the current solely in the real Marconi wire and not in its image also.

By taking the energy flow at any point above ground, we quadruple it to get the total energy flow due to the Marconi wire and its image, since doubling the field quadruples the energy but the assumed perfectly reflecting and therefore isolating earth surface, halves this amount. This treat-

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ment of course, fails, close to the antenna, between say ¼ wavelength and 10 wavelengths, but these distances are of no importance.

Stone and myself treated the Marconi wire as an unlimited number of Hertz "doublets" each carrying the same infinitesimal current element "*di*" but each one having a length proportional to the ordinate of the current curve 4-d-t at any point along a-4 in Fig. 1-C. This figure is shown for simplicity with 6 current elements, each having a value

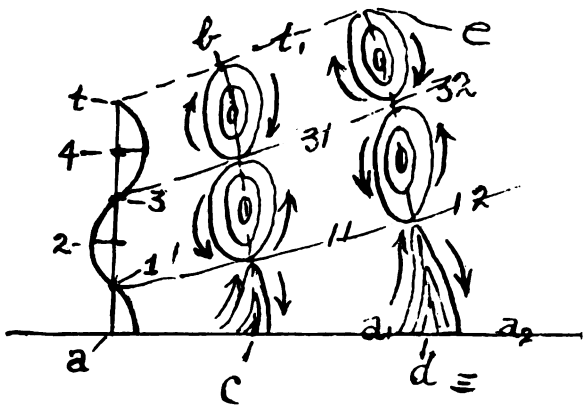

Figure 2

of ¼ of the unit current a-4 at the base of the Marconi wire.

The "length" of these currents as compared to the Marconi wire length a-t is given by the average of a-t and 1-1, 11-22, 22-33, 33-44, 44-55, 55 and the average of 55 with zero. Integrating an infinite number of these "current doublets" we had an oscillating "current sheet", composed of an infinite of oscillating undamped current elements of length from 0 to the full height of the Marconi wire.

By dividing the field produced at a distant point as a result of the action of all these "elementary" oscillators by the unit current we got the average field strength produced by a single Hertz doublet with unity current and height equal to that of the Marconi wire. This is the concept known as the "effective height" of an antenna.

Where Stone differed from recent mathematicians in treating these higher frequencies, is in the fact that he did

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not contemplate current excitation at a loop of current or voltage excitation at a loop of potential by external means. All transmitter excitation was calculated on the basis of forced continuous wave voltage at the antenna base.

In embodying his work in 1905 and 1906 on practical receivers, I discovered that on the above basis of operation a transmitting antenna had infinite resistance as well as infinite reactance at even multiples of the fundamental frequency, since the current maximum was theoretically

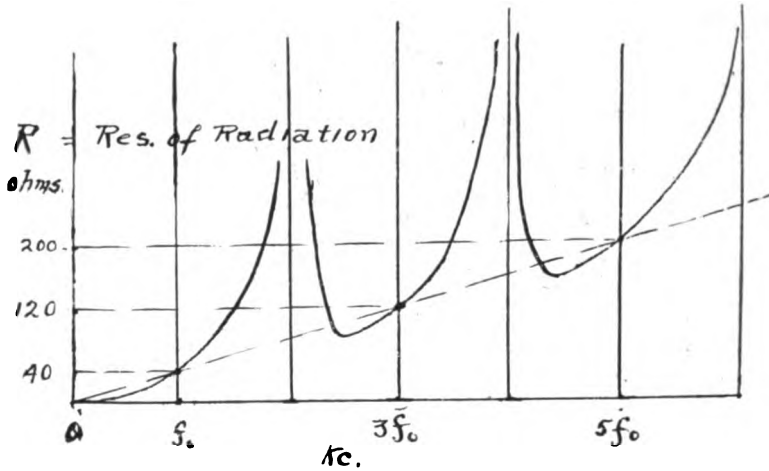

Figure 3

without limit relative to the antenna current at the base at such frequencies—if increased voltage was available to keep the base current constant. This made antenna resistance a periodic function, under the conditions above given.

When the frequency is an odd multiple of the fundamental value, the resistance equivalent of radiation becomes a simple odd multiple of that at the fundamental, e. g. 40-120-200 etc. ohms approximately, not allowing for azimuthal effects, as these were of no interest then. We had not then learned to "carom" on the electron cushion of the radio vault in order to play "celestial billiards"—in skipping oceans and lesser obstructions.

Hack's Work

Coming now to the upper multiple frequencies of the Marconi wire, such as the 5th etc. we find interesting fac-

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tors bearing on Messrs. Levin and Young's paper. Reference may be had to *F. Hack, Am. Phys. Band 14, p. 539, 1904 and Band 18, p. 634, 1905*, for details of Hack's work.

What Hack showed was that in a Marconi antenna, as shown in Fig. 2—vibrating at its quintuple "free" or unloaded frequency—we have 2 complete sets of "free" waves from the segments t-3 and 3-1 respectively and a half of sliding wave from the segment a-1. From the properties of the electromagnetic field these "contacting" waves mutually repel each other and the top ones are forced further upwards on their way outward.

What is interesting to consider is this: Since these partial but self-sustaining harmonic radiations do not merge into each other, their energy equivalents should be expressed at the antenna base by a system of purely additive scalar resistances. In other words, if the resistance equivalent of the segment 1-a which is at its fundamental or "free" frequency—is about 40 ohms, then both segments 1-3 and 3-4 add 80 ohms respectively to the radiation resistance, making a total of about 200 ohms at a distant point near the equatorial or earth plane.

This seems to sustain my point of view in the July 4, 1908 Radio Review article that with constant current at the base the resistance equivalent of radiation varies with frequency approximately as the ordinates of Fig. 3.

It may be possible to bring to light these old records with Mr. Stone's permission. These results with Prof. Howes' recent work along the same general lines as Messrs. Levin and Young have so painstakingly worked out may well clear up any discrepancies which appear in tests of these theories.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELPHONY

Issued Jan. 4, 1927—Feb. 8, 1927

By

JOHN B. BRADY

Patent Lawyer, Ouray Building, Washington, D. C.)

1,612,835—INTERMEDIATE ELECTRODE IN INCANDESCENT CATHODE TUBE—WALTER SCHOTTKY, Wurzburg, Germany. Filed Aug. 30, 1921, issued Jan. 4, 1927. Assigned to Siemens & Halske Aktiengesellschaft.

1,612,902—RADIOCONDENSER—C. V. BATES, Chicago, Ill. Filed April 16, 1925, issued Jan. 4, 1927. Assigned to Cinch Mfg. Corp.

1,612,952—RADIO RECEIVING APPARATUS—G. H. STEVENSON, New York, N. Y. Filed Aug. 24, 1921, issued Jan. 4, 1927. Assigned to Western Electric Co., Inc.

1,613,032—VARIABLE CONDENSER—C. F. GOUDY, Flushing, N. Y. Filed Oct. 30, 1924, issued Jan. 4, 1927. Assigned to Pacent Electric Co., Inc.

1,613,088—CONDENSER CONTROL FOR RADIOS—W. G. DUNN, Clarinda, Ia. Filed March 3, 1926, issued Jan. 4, 1927.

1,613,333—LOOP AERIAL—J. McW. STONE, Chicago, Ill. Filed Oct. 31, 1923, issued Jan. 4, 1927. Assigned to Operadio Corp.

1,613,503—ELECTRON TUBE—W. DAUMANN, Hamburg, Germany. Filed Jan. 16, 1923, issued Jan. 4, 1927. Assigned to C. H. F. Muller.

1,613,607—COMPENSATING NETWORK FOR CARRIER TRANSMISSION CIRCUITS—H. FLETCHER, New York, N. Y. Filed Sept. 29, 1924, issued Jan. 11, 1927. Assigned to Western Electric Co.

1,613,791—ELECTRON TUBE CONSTRUCTION—M. B. FORRESTER, Philadelphia, Pa. Filed Apr. 11, 1925, issued Jan. 11, 1927. Assigned to J. S. Timmons, Inc.

1,613,612—ELECTRON DISCHARGE DEVICE—ROBERT W. KING, New York City. Filed August 14, 1919, issued Jan. 11, 1927. Assigned to Western Electric Co.

1,613,626—SPACE DISCHARGE SYSTEM—J. J. VAN DER BIJL, New York City. Filed Oct. 30, 1918, issued Jan. 11, 1927. Assigned to Western Electric Co.

1,613,630—VACUUM TUBE—P. M. WATROUS, Jersey City, N. J. Filed Nov. 25, 1919, issued Jan. 11, 1927. Assigned to Western Electric Co.

1,613,709—ELECTRIC SPARK GAP—C. L. A. M. LEBLANC, Paris, France. Filed Apr. 18, 1923, issued Jan. 11, 1927. Assigned to Societe Anonyme Pour L'Exploitation Des Procedes Maurice LeBlanc-Vickers.

1,613,733—RADIO SIGNAL CIRCUITS—A. H. TAYLOR, Washington, D. C. Filed Jan. 7, 1925, issued Jan. 11, 1927. Assigned to Wired Radio, Inc.

1,614,113—ELECTRIC CONDENSER—P. R. FORTIN, Schenectady, N. Y. Filed May 13, 1925, issued Jan. 11, 1927. Assigned to General Electric Co.

1,614,154—ELECTRON DISCHARGE APPARATUS—O. W. PIKE, Schenectady, N. Y. Filed Mar. 2, 1925, issued Jan. 11, 1927. Assigned to General Electric Co.

1,614,428—VARIABLE CONDENSER—F. O. HARTMAN, Mansfield, O. Filed Nov. 5, 1925, issued Jan. 11, 1927. Assigned to Hartman Electrical Mfg. Co.

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1,614,458—RADIO RECEIVING SYSTEM—O. E. CURTIS, Pittsford, N. Y. Filed May 3, 1923, issued Jan. 18, 1927.

1,614,459—ELECTRICAL CONDENSER—W. T. DITCHAM, Colchester, England. Filed Apr. 19, 1923, issued Jan. 18, 1927. Assigned to Radio Corp. of America.

1,614,462—THERMIONIC DEVICE—H. M. DOWSETT, Colchester, England. Filed Nov. 12, 1920, issued Jan. 18, 1927. Assigned to Radio Corp. of America.

1,614,484—COLLAPSIBLE LOOP AERIAL—M. MATHIESEN, Chicago, Ill. Filed Aug. 1, 1925, issued Jan. 18, 1927. Assigned to Mathiesen Sandberg Co.

1,614,494—CIRCUIT ARRANGEMENT FOR GENERATING OSCILLATIONS—H. RUKOP, Berlin, Germany. Filed Dec. 27, 1922, issued Jan. 18, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie.

1,614,671—AUTOMATIC ANTENNA SYSTEM—J. H. HAMMOND, JR., Gloucester, Mass. Original filed July 9, 1920; Division filed Dec. 2, 1922, issued Jan. 18, 1927.

1,614,762—HIGH POTENTIAL ELECTRICAL APPARATUS—J. A. PROCTOR, Lexington, Mass. Filed Dec. 12, 1923, issued Jan. 18, 1927. Assigned to Wireless Specialty Apparatus Co.

1,614,902—VARIABLE CONDENSER—J. SCHOLTZ & G. W. SUNDAY, Philadelphia, Pa. Filed Feb. 13, 1925, issued Jan. 18, 1927.

1,614,990—RADIOSIGNALING ANTENNA SYSTEM—P. O. PEDERSEN, Fredericksberg, Germany. Filed July 21, 1921, issued Jan. 18, 1927.

1,615,093—HIGH VACUUM SPACE DISCHARGE DEVICE AND GAS ACCUMULATOR THEREFOR—F. S. McCULLOUGH, Wilkinsburg, Pa. Filed Sept. 17, 1923, issued Jan. 18, 1927.

1,615,224—CHOKE COIL AMPLIFICATION UNIT—EDWARD E. HILLER, Bloomfield, N. J. Filed Apr. 30, 1926, issued Jan. 25, 1927.

1,615,468—ANTIMICROPHONIC SHIELD FOR ELECTRON TUBES—LESLIE R. McDONALD, Montreal, Quebec, Canada. Filed July 6, 1926, issued Jan. 25, 1927.

1,615,645—COMBINED RADIO SENDING AND RECEIVING SYSTEM—ALEXANDER NYMAN, Wilkinsburg, Pa. Filed July 15, 1920, issued Jan. 25, 1927. Assignee Westinghouse Electric & Mfg. Co.

1,615,808—VACUUM TUBE CIRCUITS—MARIUS LATOUR, Paris, France. Filed Aug. 19, 1921, issued Jan. 25, 1927. Assignee Latour Corporation.

1,615,654—VACUUM TUBE STRUCTURE AND METHOD OF ASSEMBLING THE SAME—M. H. RYDER, of Wilkinsburg, Pa. Filed Nov. 7, 1922, issued Jan. 25, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,615,660—VACUUM TUBE OSCILLATOR—J. SLEPIAN, Swissvale, Pa. Filed Feb. 26, 1921, issued Jan. 25, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,615,712—SYSTEM OF NAVIGATION AND PLANT FOR CARRYING IT OUT—W. A. LOTH, Paris, France. Filed Feb. 27, 1925, issued Jan. 25, 1927. Assigned to Societe Industrielle des Procedes W. A. Loth.

1,615,755—LOOP ANTENNA—F. GRAUE, Washington, D. C. Filed Nov. 11, 1925, issued Jan 25, 1927. Assigned to George T. Kemp.

1,615,896—HIGH FREQUENCY SIGNALING SYSTEM—H. A. AFFEL, Brooklyn, N. Y. Filed Dec. 15, 1922, issued Feb. 1, 1927. Assigned to American Telephone and Telegraph Company.

1,615,913—RADIO DIRECTION FINDING SYSTEM—H. S. OSBORNE, Montclair, N. J. Filed Dec. 6, 1920, issued Feb. 1, 1927. Assigned to American Telephone & Telegraph Company.

1,615,958—ELECTRON DISCHARGE DEVICE—P. SCHWERIN, New York, N. Y. Filed Dec. 15, 1923, issued Feb. 1, 1927. Assigned to Western Electric Co.

1,616,010—DESTRUCTION OF THE ELECTRICAL CONDUCTIVITY OF FILMS—C. T. ULREY, East Orange, N. J. Filed Jan. 3, 1923, issued Feb. 1, 1927. Assigned to Westinghouse Lamp Co.

1,616,093—VARIABLE CONDENSER FOR RADIOCIRCUITS—F. W. STEIN, Atchison, Kan. Filed May 2, 1924, issued Feb. 1, 1927.

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1,616,176—VIBRATION ABSORBER FOR RADIO TUBES—H. A. BREMER, Chicago, Ill. Filed Nov. 7, 1925, issued Feb. 1, 1927.

1,616,214—VACUUM TUBE—J. L. BROWN, Long Beach, Cal. Filed May 5, 1922, issued Feb. 1, 1927.

1,616,139—ELECTRON DISCHARGE DEVICE—V. L. RONCI, Brooklyn, N. Y. Filed Feb. 21, 1925, issued Feb. 1, 1927. Assigned to Western Electric Co.

1,616,184—ELECTRON DISCHARGE DEVICE—T. R. GRIFFITH, Dover, N. J. Filed Oct. 29, 1923, issued Feb. 1, 1927. Assigned to Western Electric Co.

1,616,622—OSCILLATION GENERATOR WITH AUTOMATIC FREQUENCY CONTROL—J. W. HORTON, Bloomfield, N. J. Filed May 21, 1923, issued Feb. 8, 1927. Assigned to Western Electric Co.

1,616,892—DUPLEX RADIO SYSTEM—L. ESPENSCHIED and DE LOSS K. MARTIN, of Hollis, N. Y., and Orange, N. J. respectively. Filed Aug. 18, 1922, issued Feb. 8, 1927. Assigned to American Telephone & Telegraph Co.

1,616,914—ELECTRON DISCHARGE DEVICE—A. MAVROGENOS, Milwaukee, Wis. Filed Jan. 15, 1926, issued Feb. 8, 1927.

1,616,923—INTERFERENCE REDUCING MEANS FOR RADIO RECEIVING APPARATUS—R. H. RANGER, Brooklyn, N. Y. Filed Dec. 28, 1922, issued Feb. 8, 1927. Assigned to Radio Corp. of America.

1,617,023—AERIAL—J. E. MONTELIUS, Los Angeles, Cal. Filed Dec. 27, 1924, issued Feb. 8, 1927.

1,617,171—METHOD AND APPARATUS FOR THE RECTIFICATION OF ALTERNATING CURRENTS—C. G. SMITH, of Medford, Mass. Filed Oct. 8, 1920, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,172—PRODUCTION OF ELECTRICAL VARIATIONS—C. G. SMITH, of Medford, Mass. Filed Oct. 8, 1920, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,173—PRODUCTION OF ELECTRICAL ENERGY VARIATIONS—C. G. SMITH, of Medford, Mass. Filed Oct. 20, 1920, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,174—ELECTRICAL APPARATUS AND METHOD—C. G. SMITH, of Medford, Mass. Filed Apr. 25, 1921, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,175—ELECTRICAL APPARATUS—C. G. SMITH, of Medford, Mass. Filed Apr. 25, 1921, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,176—ELECTRICAL APPARATUS—C. G. SMITH, of Medford, Mass. Filed May 3, 1921, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,177—PRODUCTION OF ELECTRICAL VARIATIONS—C. G. SMITH, of Medford, Mass. Filed May 3, 1921, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,178—ELECTRICAL APPARATUS—V. BUSH, of Chelsea, Mass. Filed Oct. 27, 1921, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,179—ELECTRICAL APPARATUS—C. G. SMITH, of Medford, Mass. Filed Dec. 31, 1921, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,180—ELECTRICAL APPARATUS—C. G. SMITH, of Medford, Mass. Filed Jan. 3, 1921, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,181—ELECTRICAL APPARATUS—V. BUSH, of Medford, Mass. Filed Apr. 7, 1922, issued Feb. 8, 1927. Assigned to Raytheon Mfg. Co.

1,617,236—PORTABLE RADIO RECEIVING SET—H. CSANYI, of New York. Filed Dec. 7, 1922, issued Feb. 8, 1927.

1,617,240—METHOD FOR RADIO DIRECTIVE SIGNALING—R. A. FESSENDEN, of Chestnut Hill, Mass. Filed Jan. 28, 1922, issued Feb. 8, 1927.

1,617,242—RADIO TRANSMISSION AND RECEPTION—R. A. FESSENDEN, of Chestnut Hill, Mass. Filed Oct. 13, 1924, issued Feb. 8, 1927.

1,616,532—CONDENSER—R. C. SPRAGUE, of Quincy, Mass. Filed Sept. 18, 1925, issued Feb. 8, 1927.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED APRIL 6, 1927

Transferred to Member Grade

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	New York 181 Sullivan St.,	Harkness, Kenneth
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	Cleveland, 345 The Arcade,	Browne, T. C.
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	San Francisco, USS Pennsylvania,	Fraser, Graham
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	Hingham, Box 414	Davidson, Martin
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	Detroit, 1507 Cass Ave.,	Swartz, Arthur R.
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	Glen Rock, 102 Glen Ave.,	Bailey, Ralph G.
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	Brooklyn, 148-83rd St.,	Fiala, Anthony
	Brooklyn, 321 Court St.,	Goodwin, Rob't. F.
	Brooklyn, 499 E. 18th St.,	Hewlett, Oscar H.
	Brooklyn, 1761-68th St.,	Moskowitz, Morris
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	Flushing, 183 Union St.,	Crombie, Kenneth F.
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	New York, 11 Essex St.,	Chernow, Morris
	New York, 328 E. 180th St.,	Geloso, John
	New York, 3030 Heath Ave.,	Hedges, Ellsworth J.
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	New York, 111 East 42nd St.,	Tannen, Sydney S.
	New York, 50 West Tremont Ave.,	Wellman, Wm. R.
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	Sea Gate, 3715 Atlantic Ave.,	Hemberger, E. F. Jr.
	Schenectady, 13 Ninth St.,	Dome, Robert B.
	Schenectady, 8 Grosvenor Square,	Douthit, M. L.
	Syracuse, 308 James St.,	Topp, Rob't. J.
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	Cleveland, 2134 Lee Road,	Baier, Henry J.
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	Cleveland, 8926 Euclid Ave.,	Hadden, Edwin R.
	Cleveland, 1944 E. 66th St.,	Harrington, J. G.
	Cleveland, 2187 Cummington Road,	Hills, Humphrey
	Cleveland, 2200 Prospect Ave.,	Hunt, Louis W.
	Cleveland, 1760 E. 19th St.,	Jackson, Robert
	Cleveland, 2203 Lee Road,	Kaestle, Albert C.
	Cleveland, 1431 Addison Road,	King, H. M.
	Cleveland, 2560 E. 130th St.,	Kintner, Dean S.
	Cleveland, 1430 W. 9th St.,	Lavo, John H.
	Cleveland, 1340 Ansel Road,	Logan, Frank T.
	Cleveland, East 131st St.,	Lundeen, E. F.
	Cleveland, 1581 Elderwood Ave.,	Lyman, Lawrence L.
	Cleveland, 2979 Hampshire Road	Maclaren, Wallace
	Cleveland, 9207 Birchdale Ave.	Maddy, Walter E.
	Cleveland, 2051 Abington Road,	Manley, Edward
	Cleveland, 829 E. 88th St.,	Morris, C. R.
	Cleveland, 246 East 131st St.,	Obenchain, F. P.
	Cleveland, 1911 Forestdale Ave.,	O'Neill, T. G.
	Cleveland, 1468 East 111th St.,	Peoples, A. G.
	Cleveland, 2-26 Union Trust Bldg.,	Plaisted, Ross J.
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	Clev. Heights, 2096 Surry Road,	Wall, John, Jr.
	Cleveland, 1599 St. Clair Ave.,	Wolbarst, Herman
	East Cleveland, 14805 Ardenall Ave.,	Shipman, Chas. H
	East Cleveland, 1880 Wymore Ave.,	Siddall, Richard
Pennsylvania,	Lancaster, Edison Elec. Co.,	Stauffer, J. Luke
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	Pittsburgh, 969 Washington Blvd.,	Lightfoot, Thos. C.
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	Punxsutawney, 206 Greenwood Ave.,	Bowser, Jacob W.
	Turtle Creek, 409 Albert St.,	Ralston, Ben C.
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Tennessee,	Nashville, Station WSM,	Montgomery, L. H. Jr.
Texas,	Dallas, 4521 Bryan St.,	McGalin, A. G.
	Dallas, 522 Cumberland St.,	Robinson, Harrold J.
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	Richmond Highlands	Walker, Rob't M.
West Virginia,	Wheeling, 89 Indiana St.	Pence, S. E.
Wisconsin,	Madison, 1326 Morrison St.,	Holman, Henry
	Milwaukee, 1305 Wisconsin Ave.,	Lighty, Russell D.
	Milwaukee, 175 Farwell Ave.,	Thomas, Morris W.

Associates Outside the U. S.

Australia,	Victoria, 511 Havelock St.,	McCullough, G. R.
Canada,	Montreal, 11 St. Sacrament St.,	Rogers, H. A.
	Ontario, Hydro,	Ferguson, A. F.
	Ontario, Napanee,	Caton, Wm. A.
	Toronto, 44 Wheeler Ave.,	Heywood, J. L.

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	Toronto, 476 Markham St.,	McMillan, D. B.
	Toronto, 12 Badgerow Ave.,	Whitmore, W. R.
	Toronto, 21 McKenzie Crescent,	Wilson, Clinton
	Winnipeg, 100 Lenore St.,	Mills, G. H.
Canal Zone,	Corozal, 10th Sig. Co.,	Blake, Garland C.
China,	Shanghai, 12 Nanking Road,	Hsu, Philip H.
England,	Birkenhead, 138 Bedford Road,	Miller, Wm. H.
	London, 34 Deans Road,	Dyer, H. John
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	Southampton, "Tregarthen,"	Rowland, Edgar S.
	Wimbledown, 23 Park Road,	Williams, Eric
Irish Free State,	Dublin, 131 Emmet Road,	Farrell H. J.
New Zealand,	Dunedin, Box 261,	Cunningham, John
Ter. Hawaii,	Honolulu, P. O. Box 3054,	Schaeffer, Carl T.

Juniors Elected

Colorado,	Denver, 611 Marion St.,	Lasky, Philip G.
Wisconsin,	Madison, 1326 Morrison St.,	Holman, Henry
Canada,	Toronto, 15 Lennox Ave.,	Hird, A. R.

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PROCEEDINGS OF

The Institute of Radio Engineers

VOLUME 15 JUNE, 1927 NUMBER 6

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GENERAL INFORMATION

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PROCEEDINGS OF

The Institute of Radio Engineers

VOLUME 15 JUNE, 1927 NUMBER 6

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GENERAL INFORMATION

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INSTITUTE ACTIVITIES

Preliminary Reports of I. R. E. Standardization Subcommittees

The preliminary drafts of reports of three of the Subcommittees of the Standardization Committee have been compiled and are now available in printed form for distribution to interested members and non-members of the Institute.

These reports have been submitted by the Subcommittee on Vacuum Tubes, Receiving Sets and Electro-Acoustic Devices.

It is desired that a copy of these reports be placed in the hands of every person interested in radio standardization in order that comments and criticisms may be made before the reports have been acted upon by the Standardization Committee.

A copy of the reports will be mailed promptly upon application to the headquarters of the Institute.

May Meeting of Board of Direction

At the meeting of the Board of Direction of the Institute held on May 4, 1927 the following were present: Dr. Ralph Bown, President; Dr. A. N. Goldsmith, Secretary; W. F. Hubley, Treasurer; Donald McNicol, Junior Past President; Melville Eastham, L. A. Hazeltine, R. A. Heising, J. V. L. Hogan, R. H. Manson, R. H. Marriott, L. E. Whittemore and J. M. Clayton, Assistant Secretary.

The following Associate members were transferred to the grade of Member: F. Cheyney Beekley, Donald G. Little and J. A. J. Cooper.

Trinidad Matres was elected to the grade of Member.

One hundred and seventeen Associates and twenty-five Juniors were elected.

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New York Meeting of the Institute

At the May meeting of the Institute held in the Engineering Societies Building, 33 West 39th Street, New York, on May 4, 1927 two papers were presented.

The first one, by Dr. Irving Wolff and A. Ringel, on "Loud Speaker Testing Methods" was presented by Dr. Wolff. This paper was illustrated by lantern slides. Copies were also available in preprint form.

The second paper was "On Quality of Speech and Music", presented by J. B. Kelly. Demonstrations to the effect of dropping frequencies from voice and music were given.

The papers were discussed by Messrs. Bown, Herbst, Frederick, Maloff, Fletcher, Vreeland, Furness, Minton, Cutting, Austin, Ringel, Wolff and others.

The attendance at this meeting was over three hundred and fifty.

No July and August Institute Meetings

There will be no meetings of the Institute during the months of July and August. The Institute offices will be open as usual, however, for the transaction of Institute business.

News of the Sections

Philadelphia Section

The March meeting of the Philadelphia Section was held on the 25th of the month in the Bartol Laboratories, Philadelphia. J. C. Van Horn was the presiding officer.

A paper was presented by W. W. Knight on, "The Rejuvenation of Composite Radio Broadcasting Stations". This paper was discussed by many of the members.

The attendance at this meeting was over one hundred.

The April meeting of the Philadelphia Section was held in the Bartol Laboratories on April 22nd. J. C. Van Horn was the presiding officer.

Roger W. Barrington, Jr. read a paper, "Development of the Radio Receiver". This paper was discussed by R. S. Milne, R. L. Snyder and others.

The attendance was one hundred and twenty-five.

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Following the meeting an inspection of the Bartol Laboratories was made by the members.

The next meeting of the Philadelphia Section will be held on May 27th in the Bartol Laboratories. Professor Charles Weyl of the University of Pennsylvania will deliver a paper entitled, "Reproduction of Sound".

Cleveland Section

The Cleveland Section held a meeting in the Case School of Applied Science on April 22nd. Ralph E. Farnham presided.

Two papers were presented. The first on, "Transmission of Pictures by Wire" was by R. S. Breese of the American Telephone & Telegraph Company.

The second by John A. Victoreen of the Victoreen Research Laboratories was on, "Photo Electric Cells".

The two papers were discussed by the members present. There were fifty-nine members in attendance.

The next meeting of the Cleveland Section will be held on May 26th in the Case School of Applied Science at which time W. C. Blackburn of the Western Electric Company will deliver a paper on "High Quality Reproduction".

Canadian Section

On May 4th a meeting of the Canadian Section was held in the Electrical Building of the University of Toronto. D. Hepburn presided.

A paper on "Electrolytic Condensers for 'A' Current Filters" was read by C. I. Soucy. This paper was presented by E. F. Andrews before the Chicago Section on February 25, 1927.

The result of the annual general meeting of the Section in the election of new officers of the Section is as follows: Honorary Chairman, Professor T. R. Rosebrugh; Chairman, A. M. Patience; Vice-Chairman, C. I. Soucy; Secretary-Treasurer, C. C. Meredith and Assistant Secretary, V. G. Smith.

Following this meeting ten manufacturers of radio equipment had arranged for a display of their new apparatus at the meeting place.

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The Canadian Section will hold its next meeting on September 1st, 1927.

Seattle Section

A meeting of the Seattle Section was held in the Club Room of the Telephone Building, Seattle, on April 9th. W. A. Kleist presided.

Two papers were read. The first was entitled, "Copper Oxide Rectifiers" and was presented by J. A. Burleigh. The second paper was by J. J. Ritter on, "A New Type of Electrolytic Condenser".

These papers were discussed by Messrs. Burleigh, Tolmie, Anderson, Syllvester, Libby and others.

The Chairman announced the appointment of the following Standing Committees of the Section: Meetings and Papers Committee—J. R. Tolmie (Chairman), H. E. Renfro, J. W. Greig and H. F. Mason. Membership Committee—C. E. Williams (Chairman), J. A. Burleigh and M. L. Wooley.

Connecticut Valley Section

On May 6, 1927 a meeting of the Connecticut Valley Section was held in the auditorium of the Hartford Electric Light Company, Hartford, Conn. Dr. W. G. Cady was the presiding officer.

The meeting was addressed by D. G. Little of the Westinghouse Electric and Manufacturing Company on, "Some Experiences as a Radio Engineer" in which the speaker described the design of recent high power short wave radio equipment.

The discussion was participated in by Dr. Cady, Dr. Van Dyke and Messrs. Little, Hull, Kruse, Bourne, French, Brackett and others.

Rochester Section

The Rochester Section held its April meeting on the 15th of the month in the Sagamore Hotel, Rochester. H. J. Klumb presided.

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The meeting was addressed by Charles Bartlett who gave a talk on "Measuring Instruments Used in Radio".

Forty members attended the meeting.

The next meeting of the Rochester Section will be held on May 20th in the Sagamore Hotel. B. Olney will read a paper on "Acoustics".

Los Angeles Section

On April 18th the Los Angeles Section held a meeting in the Los Angeles Elks Club. Don Wallace presided.

Harry Fore of the Philadelphia Storage Battery Company presented a talk on "Socket Power Devices, Their Construction and Operation".

There were thirty-five members present.

Washington Section

The Washington Section held a meeting on April 13th in Harvey's Restaurant, 11th and Pennsylvania Avenue, N. W., Washington. Dr. J. H. Dellinger presided.

Dr. L. P. Wheeler of the Naval Research Laboratory delivered a talk on "The Establishment of Standard Radio Frequencies".

It was decided to hold future meetings of the Washington Section on the second Thursday of each month instead of the second Wednesday.

The next meeting of the Section will be held on May 12th at which time Commander S. C. Hooper, U. S. N. will deliver a paper on "The Mission and Scope of Naval Radio". This meeting will be held in Harvey's Restaurant.

Proposed New Sections

Activity in the establishment of new Sections of the Institute continues. Correspondence looking to the formations of Sections in Milwaukee, Wisconsin; Atlanta, Georgia and Buffalo, New York is being held, in addition to the Section formation activities previously mentioned in the Proceedings.

In Milwaukee two organization meetings have been held. D. S. W. Kelly has been elected chairman of the temporary

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organization and S. Snead has been elected temporary Secretary.

In Atlanta, the preliminary organization work is being sponsored by The Atlanta Association of Radio Engineers, Major Walter Van Nostrand being the chairman of the Committee on Organization of an Institute Section. One organization meeting has been held, and it is contemplated that additional meetings will be held in the near future.

The Buffalo work has been initiated by C. J. Porter. The first organization meeting will be held in the near future.

Past Numbers of the PROCEEDINGS

Back copies of the PROCEEDINGS, except as listed below, are available from the Institute. From Vol 1 to Vol. 14, inclusive, these copies sell at $1.50 each. Copies of Vol. 15 can be supplied at $1.00 per number. In both of these cases members of the Institute are entitled to a discount of 25 per cent.

In the case of several of the numbers the stock of back issues is quite small. Additional copies will *not* be available when the present supply is exhausted.

As time goes on these back numbers of the PROCEEDINGS become increasingly valuable. Not only are the most important elements in the history of the engineering side of radio presented in these older issues, but practically no important phase of the radio arts or science can be named which has not been considered in the PROCEEDINGS at one time or another.

Members of the Institute who contemplate desiring to complete their files of the PROCEEDINGS should secure these copies before the stock is exhausted.

Copies listed as follows *cannot* be supplied:

Vol. 1, No. 2; Vol. 2, Nos. 3 & 4; Vol. 3, Nos. 1, 2 & 3; Vol. 4, Nos. 1 & 2; Vol. 5, No. 1; Vol. 6, No. 5; Vol. 7, Nos. 3 & 5; Vol. 9, No. 3; Vol. 11, No. 6; Vol. 12, Nos. 1, 2 & 3; Vol. 13, Nos. 1 & 4.

Bound volumes from Vol. 5 (1917) to volume No. 16 (1926) are available at $11.00 per volume, with the usual 25 per cent. discount to members.

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Fourteen Year PROCEEDINGS Index

Work is practically completed on a very comprehensive Index of the PROCEEDINGS covering the past fourteen years. This Index includes not only a list of papers and authors, but also a cross-index by means of which papers on related subjects can be located quickly.

When available a copy of the Index will be sent free of charge to each member of the Institute. The price to non-members will be one dollar per copy.

Committee Work

SECTIONAL COMMITTEE ON RADIO, A. E. S. C.

The following is a report of the activities of the Technical Committees of the Sectional Committee on Radio, A. E. S. C., for March, 1927:

Committee on Transmitting and Receiving Sets and Installations

This Committee has no activities to report during the month.

Committee on Component Parts and Wiring

The Subcommittee of this Committee is active and will report to the main Committee at its next meeting in May.

Committee on Electro-Acoustic Devices

No March meeting was held by this Committee. A meeting will be held at an early date, however, at which time the electro-acoustic standards as proposed by the National Electrical Manufacturers Association in its Year Book, and by the Subcommittee of the Institute of Radio Engineers, will be available for consideration and study.

Committee on Vacuum Tubes

Material dealing with standardization of the dimensions of vacuum tube bases is available to this Committee and will be considered by the Committee at a forthcoming meeting.

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Committee on Power Supply and Outside Plant

No activities of this Committee were reported for the month of March.

I. R. E. Subcommittee on Vacuum Tubes

At its meeting on May 6th, the Subcommittee on Vacuum Tubes completed its work, and has submitted its report to the Committee on Standardization of the Institute (this report will be found in the preliminary drafts of reports of the three Subcommittees which is available on request). The subject assigned to the Subcommittee was the recommendation of suitable methods for testing vacuum tubes. In addition to making selections from the various methods already in common use for measuring the characteristics of vacuum tubes, and specifying the conditions of test, two matters have been brought forward which had not been generally known in the art and which are especially interesting.

The *emission characteristic curve,* showing the relation of filament emission current to power which heats the filament, cannot usually be determined experimentally up to normal filament power without damaging the vacuum tube. A method of plotting experimental data, taken at reduced power, on a special "power-emission chart" results in a straight line which can be readily extended to normal power. The development of this chart is due to Dr. C. Davisson of the Bell Telephone Laboratories, who, together with M. J. Kelly of the Subcommittee, plans to present a paper on the subject in the fall.

The procedure heretofore commonly followed in measuring the *capacity between two electrodes of a vacuum tube* was to leave the third electrode on open circuit, or floating in potential. It was found that this procedure gave misleading results; it did not measure the direct capacity between the two electrodes (which is the important thing), but rather a combination of all three direct capacities present in a vacuum tube. Several methods of measuring direct capacity have been devised and studied under the direction of members of the Subcommittee. These have proven so satisfactory that it is recommended that all measurements of capacity with one electrode floating be aban-

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doned in favor of direct capacity measurements. It is hoped to have several short papers on this subject prepared for presentation to the Institute.

The notation to be employed in connection with vacuum tubes has long been a bone of contention, as indicated, for example, by the recent presentation before the Institute of a paper on that subject by Professor Chaffee. The Subcommittee has studied the question and has prepared definite recommendations. The recommendations that received the most attention were current practice and convenience in printing and typing. It was found possible to recommend a system which satisfied both requirements and which, in particular, completely avoided special character signs.

I. R. E. Subcommittee on Receiving Sets

The Subcommittee on Receiving Sets held an all-day meeting on May 5th, in order that a tentative report on its work to date could be presented to the Institute for printing. The standardization of tests on radio receiving sets opened such a large field, that the subcommittee decided to restrict its scope for the present to broadcast receiving sets. Furthermore, detailed consideration has so far been given only to over-all tests; but it is planned to include in the final report tests on the parts which make up a complete receiving set, such as radio-frequency and audio-frequency amplifying transformers.

The essential technical properties of a broadcast receiving set are sensitivity, selectivity, and fidelity. The term "fidelity" has been recommended for adoption to take the place of the very general word "quality" and to signify the accuracy of reproduction at the output terminals of the modulation of the received wave. All three properties are to be determined by impressing a simple modulated radio-frequency voltage in the antenna circuit (using the self-contained antenna, if the set is so provided—otherwise a standardized artificial or real antenna circuit), and reading the output into a non-inductive resistor of proper resistance connected in place of a loud-speaker.

To determine *sensitivity* the input is adjusted to give a so-called "normal output" of 0.05 watt at a series of wave frequencies within the broadcast band. The input voltage

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is reduced to equivalent radio field intensity, and this is plotted on a graph against wave frequency. Values of modulation frequency (400 cycles) and percentage (30%) have been tentatively set.

To determine *selectivity*, the receiving set is tuned to a certain "standard test frequency" and the radio frequency input is varied simultaneously in voltage and in frequency so as to maintain a so-called "interference output" equal to one ten-thousandth of the "normal output" (that is equal to 0.00050 watt). Equivalent radio field intensity is plotted against carrier frequency. The test is then repeated for the other "standard test frequencies", these being tentatively set at 600, 800, 1000, 1200, 1400 kilocycles.

To determine *fidelity*, the receiving set and the input are both tuned to a standard test frequency and the modulation frequency is varied from 40 to 10,000 cycles, the input voltage being held constant at the value giving normal output at 4000 cycles modulation, and the modulation being 30% throughout. The relative voltage at the output terminals is measured and plotted against frequency. The test is then repeated for the other standard set frequencies.

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SHORT-WAVE COMMERCIAL LONG-DISTANCE COMMUNICATION*

By

H. E. HALLBORG, L. A. BRIGGS, AND C. W. HANSELL

I—HISTORY AND DEVELOPMENT

The engineers of the Radio Corporation of America followed with keen interest, the short wave experiments on this side of the Atlantic conducted during the winter of 1921-1922, as well as the contemporaneous European developments in this ultra high-frequency field. This interest was further stimulated during the early short wave re-broadcasting experiments of the General Electric and Westinghouse Companies. Many measurements were made during this period with the purpose of determining whether these short wave signals would be suitable for dependable long distances telegraph communication. The results of these early rebroadcast observations were, in general, not encouraging. They indicated wide signal variations with light and darkness, and extremely bad fading.

It is not surprising in view of these early observations that the first short wave telegraph transmitter constructed by the Radio Corporation should have been designed for experimental radio relay service. This transmitter was built early in 1923 at the Transoceanic Receiving Station at Belfast, Maine. It was hoped that this installation would provide the means for studying the effectiveness of short waves for the relaying of the incoming trans-Atlantic long wave signals supplementing the land lines from Belfast, Maine to New York City, a distance of about 400 miles. A general view of this experimental short wave radio relay transmitter is shown in Figure 1.

"2XAO" the Belfast relay transmitter proved more valuable in the study of propagation phenomena, and in pointing the way to many important transmitter developments

*Received by the Editor April 4, 1927.

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than as a radio relay. The art was not sufficiently advanced at that time to bear the heavy responsibility of commercial trans-Atlantic relaying.

In the meantime, the Radio Corporation had contracted with its manufacturing companies for the construction of a

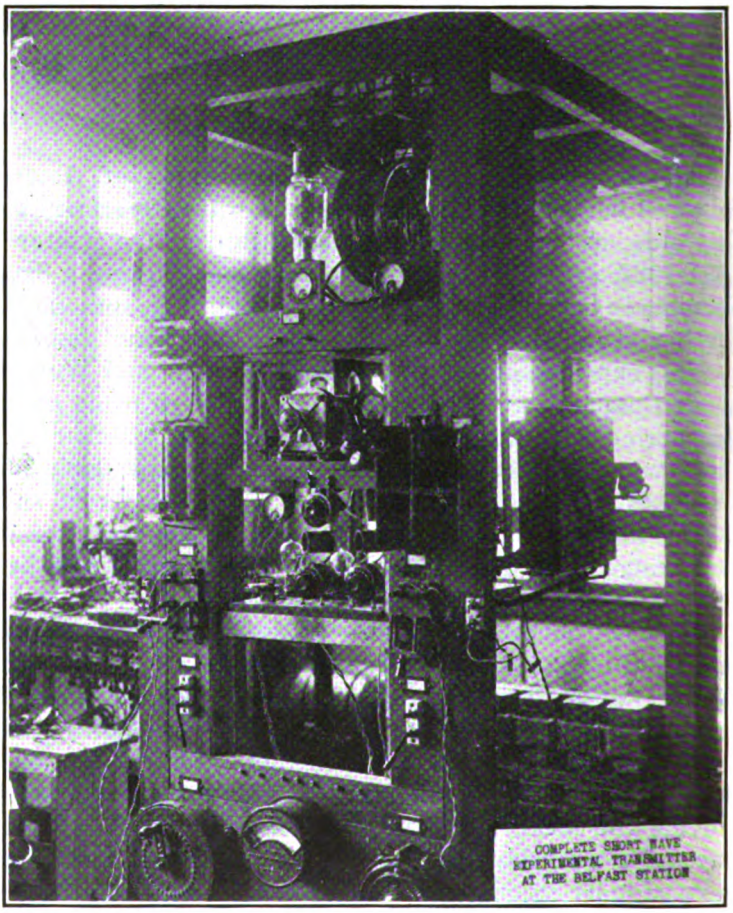

Figure 1—Experimental short wave radio relay transmitter 2XAO—Belfast, Maine.

number of experimental short wave telegraph transmitters. The first of these was installed in August 1923 at the high power trans-Atlantic Station at Tuckerton, New Jersey. It operated under the call letters "WGH" on a wavelength of 103 meters. Its first commercial use was on the evening of

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September 11, 1923 when a ringside account of the Firpo-Wills fight was sent to Buenos Aires by this transmitter.

"WGH" represents the first serious attempt by the Radio Corporation at long distance communication via short wave channels. It was used as an auxiliary to the 200-kilowatt alternators, serving intermittently for communication to Berlin, Paris, and to Buenos Aires. It was operated regu-

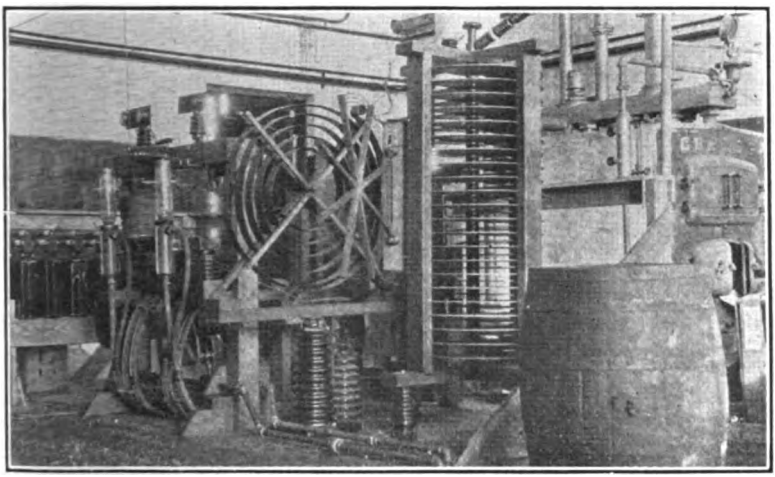

Figure 2—Experimental 103-meter short wave transmitter WGH, Tuckerton, N. J.

larly but at comparatively slow speeds for a period of several months during the hours of darkness.

A view of the original "WGH" installation is shown in Figure 2. It was of the self-excited type using two water-cooled tubes in a Hartley circuit. The self-excited circuit, due to its susceptibility to frequency fluctuations, soon gave way to a master oscillator-power amplifier type of circuit. A push-pull amplification system was adopted throughout. The two water-cooled tubes functioned as a push-pull power amplifier. The addition of the master oscillator proved of considerable value in stabilising transmission and increasing traffic speeds.

During the spring of 1924 the "WGH" installation was followed by the distribution of 5 additional experimental short wave transmitters to the various high power stations of the Radio Corporation world network. These units were

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originally of the self-excited type similar to "WGH"; but were subsequently converted to master oscillator power amplifier types and eventually fitted with piezo-electric crystal control.

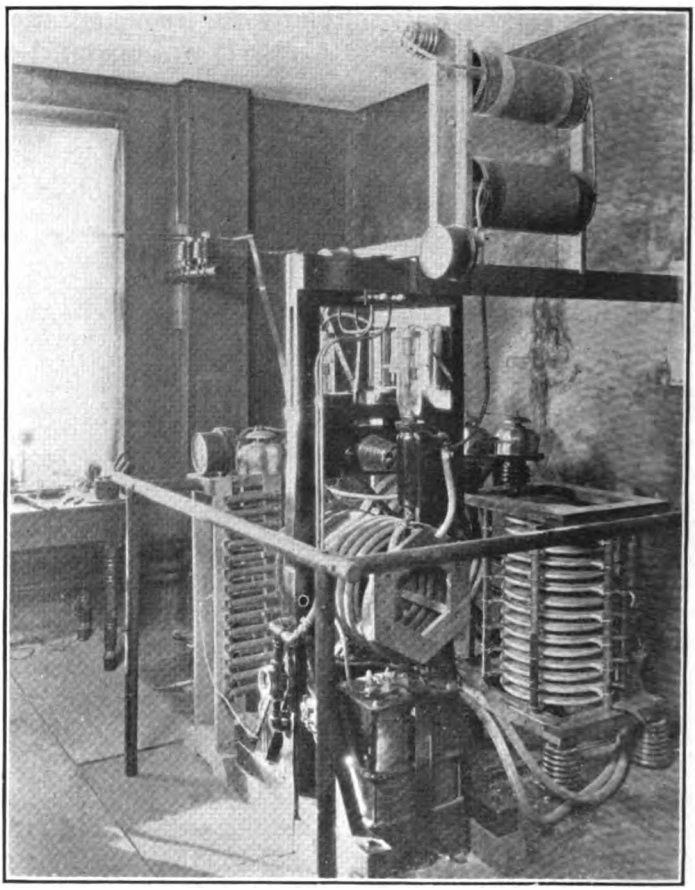

Figure 3—Experimental 95-meter short wave transmitter—KEL, Bolinas, California.

The call letters, distribution and operating wavelengths of these five transmitters were as follows:

KEL —95 meters—Bolinas, California,
KIO —90 meters—Kahuku, Hawaii,
WIR —74 meters—New Brunswick, New Jersey,
WQN—51.5 meters—Rocky Point, Long Island,
WIZ —43 meters—New Brunswick, New Jersey.

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Figure 3 is a photograph of "KEL" the 95 meter installation at Bolinas, California, which may be considered typical of this group of experimental transmitters. The master oscillator and intermediate amplifier stages which were subsequently added as shown in Figure 4. In Figure

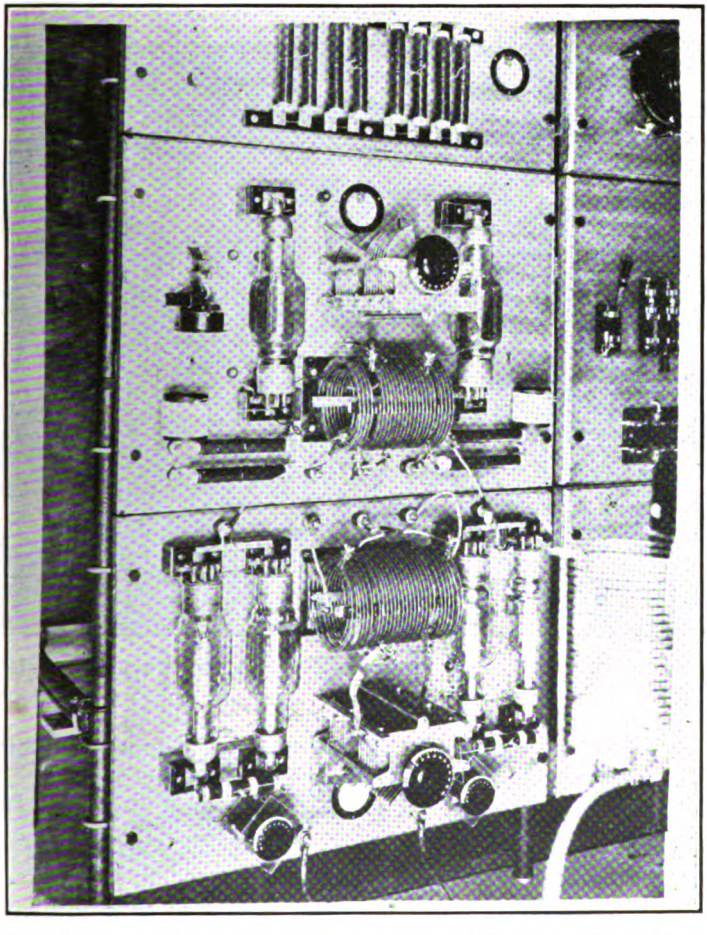

Figure 4—Experimental 95-meter master oscillator and intermediate amplifier units of short wave transmitter—KEL, Bolinas, California.

5 is shown the complete installation of the 90 meter transmitter "KIO" at Kahuku, Hawaii. This figure shows the power amplifier and intermediate amplifier panels the various control panels as well as the main rectifier for sup-

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plying 10,000 volts d. c. and the intermediate rectifier which supplied 2,000 volts d. c.

In Figure 6, is shown the 95 meter vertical antenna at Bolinas, California. This antenna was energized by means of a transmission line 875 feet long.

It is of interest to observe that of the above group of five short wave experimental transmitters, only one is still

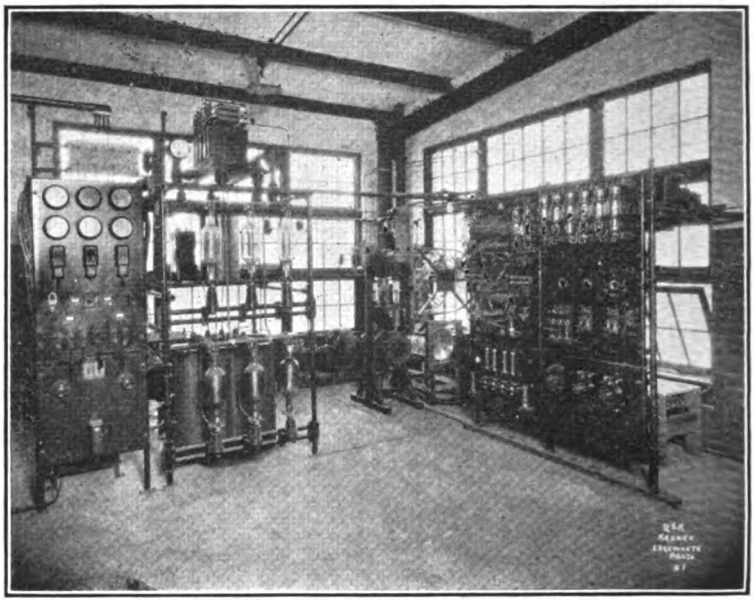

Figure 5—Complete 90-meter experimental short wave installation—KIO, Kahuku, Hawaii.

in operation on its originally assigned wavelength namely—"WIZ" 43 meters at New Brunswick, New Jersey. This transmitter operates as a crystal-controlled master oscillator, power amplifier unit directly energizing a long vertical harmonic antenna without the medium of a transmission line. It is used as a supplementary means of communication during the hours of darkness to Berlin, Paris, and Buenos Aires. "WIZ" has been one of the most effective of the Radio Corporation's experimental short wave transmitters.

Early in 1925 the five original installations above described were followed by two others:

"WQO"—35 meters—Rocky Point, Long Island,

"WIK" —22 meters—New Brunswick, New Jersey.

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"WQO" was subsequently reconverted in a reconstruction program undertaken at Rocky Point, Long Island. "WIK" was transferred from New Brunswick to Rocky Point. An intensive short wave experimental program with a view to centralizing short wave transmitter development

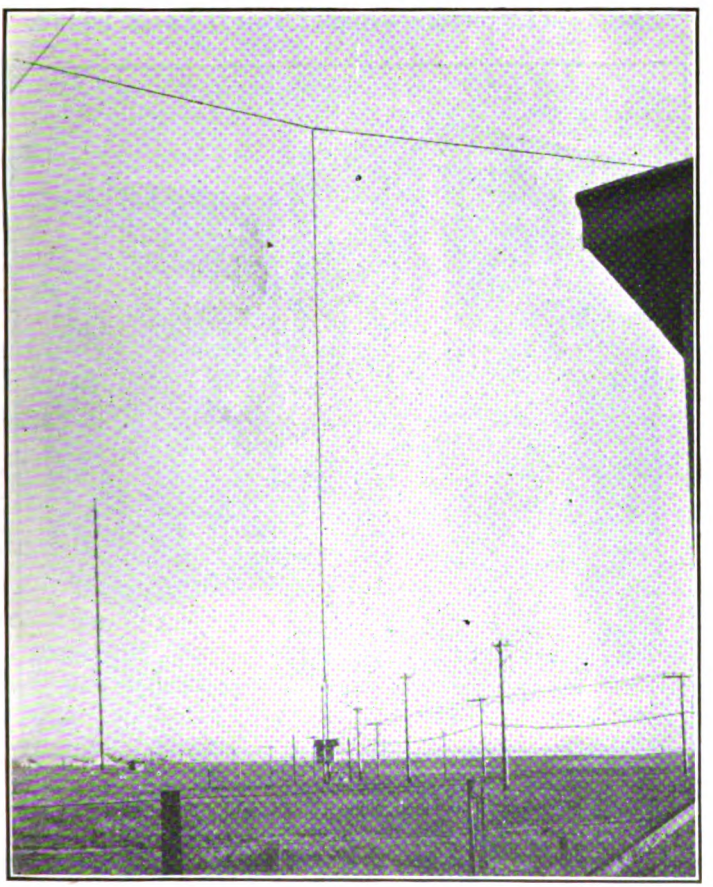

Figure 6—95-meter, third harmonic, vertical antenna and transmission line—Bolinas, California.

at Rocky Point soon resulted in the installation of three additional short wave transmitters of still lower wavelengths. These units embody all the essential features of design dictated by experience with the earlier sets, such as adequate shielding, crystal control, elimination of losses in the cooling water of the power amplifier tubes, elimination

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of interstage amplifier reactions, and improvement of tone quality.

The call letters, location and operating wavelengths of these three transmitters are as follows:

"WLL"—16.6 meters—Rocky Point, Long Island,
"2XT" —16 meters—Rocky Point, Long Island,
"2XS" —14.9 meters—Rocky Point, Long Island.

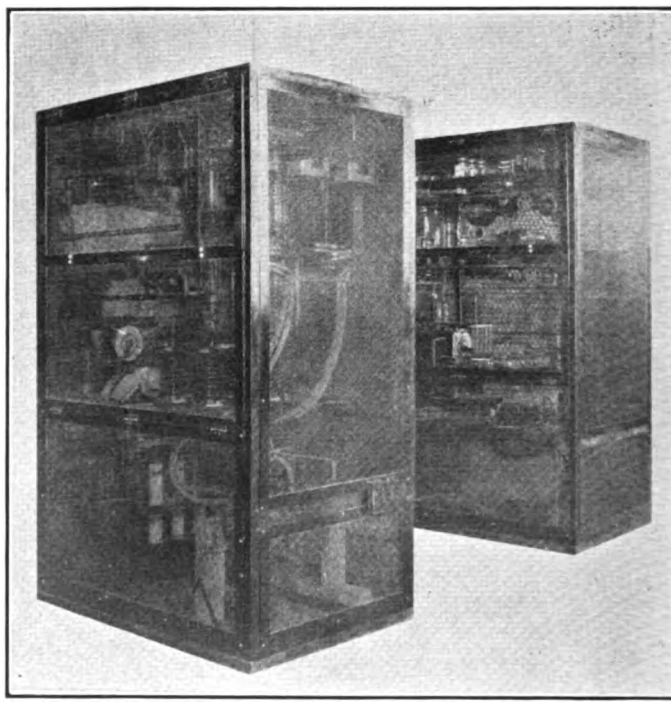

Figure 7—Short wave transmitter 2XT—16-meter, installed at Rocky Point, L. I.

This group of short wave transmitters is utilized in the Radio Corporation's South American service to Rio de Janeiro and Buenos Aires. Most interesting from a transmission viewpoint is that they are most effective during the hours of complete daylight, namely from about 6 A. M. to 6 P. M. This is shown more graphically in traffic charts in another portion of this paper.

In Figure 7, is shown a view of one of these transmitters installed for the above service at the Radio Corporation's Radio Central at Rocky Point, Long Island.

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The Radio Corporation's short wave experimental transmitters with antenna inputs varying from 3-kw. to 20-kw. have in general been used as auxiliaries to the 200-kw. long wave alternators. A summary of these various short wave installations, their call letters, wavelengths, and the service to which each was assigned, is as follows:

2XAO—Belfast, Maine—60 to 110 meters—Relay Experimental
WGH—Tuckerton, New Jersey—103 meters—to Berlin, Paris & Buenos Aires
KEL —Bolinas, California—95 meters—to Honolulu & Japan
KIO —Kahuku, Tombouy of Hawaii—90 meters—to San Francisco & Japan
WIR —New Brunswick, New Jersey—74 meters—to London, Paris & Berlin
WQN—Rocky Point, L. I., 51.5 meters—to Berlin & Paris.
WIZ —New Brunswick, New Jersey—43 meters—to Buenos Aires, Berlin & Paris
WQO —Rocky Point, Long Island—35 meters—to Berlin & Paris
IXR —Manila, Philippine Islands—30 meters—to Honolulu & San Francisco
KEL —Bolinas, California—29.3 meters—to Honolulu & Java
HJG —Bogota, Colombia—22 meters—to Central America & New York
WIK —Rocky Point, Long Island—21.5 meters—to Buenos Aires, Berlin & Paris
WLL—Rocky Point, Long Island—16.6 meters—to Rio de Janeiro & Buenos Aires
2XT —Rocky Point, Long Island—16 meters—to Rio de Janeiro & Buenos Aires
2XS —Rocky Point, Long Island—14.9 meters—Rio de Janeiro & Buenos Aires
KEL —Bolinas, California—14.1 meters—to Manila & Java

A definite trend to shorter wavelengths is indicated in the above development.

A record of the performances of these experimental transmitters on commercial traffic is believed to be a topic of general interest. This phase of the subject is considered in Section II.

II—Commercial Traffic Operations

It was realized very early in the development of short wave operation that the best test of effectiveness and comparative values of various wavelengths and types of transmitters was actual operation of the transmitter in the handling of commercial traffic. Consequently the RCA short wave transmitter instead of undergoing prolonged periods of test observation, were utilized immediately for commercial service as auxiliaries to the standard high power transmitters. The very high standards of reliability and accuracy required and maintained in the movement of commercial

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traffic provided a most severe test of short wave operations. Readability was judged not by mere audibility or even by intelligibility of signals over brief or irregular periods but by the exacting demands of comparison with established communication channels operating simultaneously with the short wave transmitters between the same two terminals.

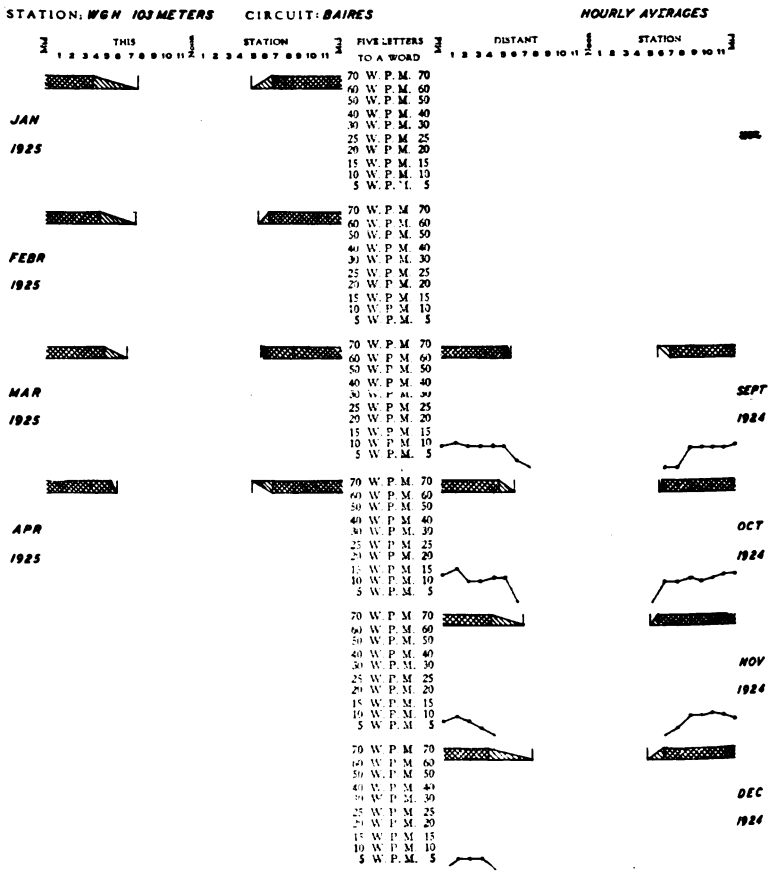

Figure 8—Commercial reception at Tuckerton—WGH, 103 meters, at Buenos Aires, September, 1924, to April, 1925.

Further improvements and refinements of the transmitters were guided by the results obtained during such service.

All the traffic data accumulated to date must be studied with the fact in mind that apparent results may have been affected to a greater or less degree by numerous alterations

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made in the transmitters from day to day. Nevertheless it seems well worth while to record the results which have been obtained and thereby furnish a basis for further and more reliable statistics in the future.

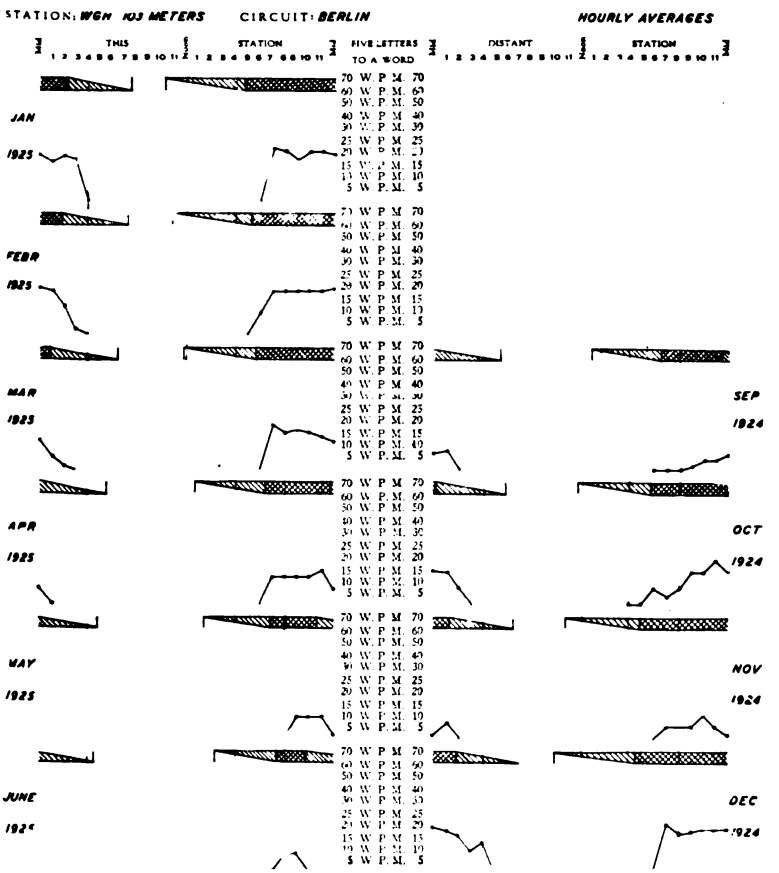

Figure 9—Commercial reception of Tuckerton—WGH, 103 meters at Berlin, September, 1924 to June, 1925.

The primary objective in all of the RCA Atlantic coast short wave activities has been to secure improved and reliable service between New York and South America. All the traffic handled and data obtained by operation between New York and Europe has been decidedly a secondary consideration. This is mentioned to explain the general trend of transmitter development and why certain transmitters and

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wavelengths with which seemingly excellent results to Europe were being obtained were abandoned in order to use available apparatus for further and more important developments in the South American services.

The first RCA short wave transmitter used for commer-

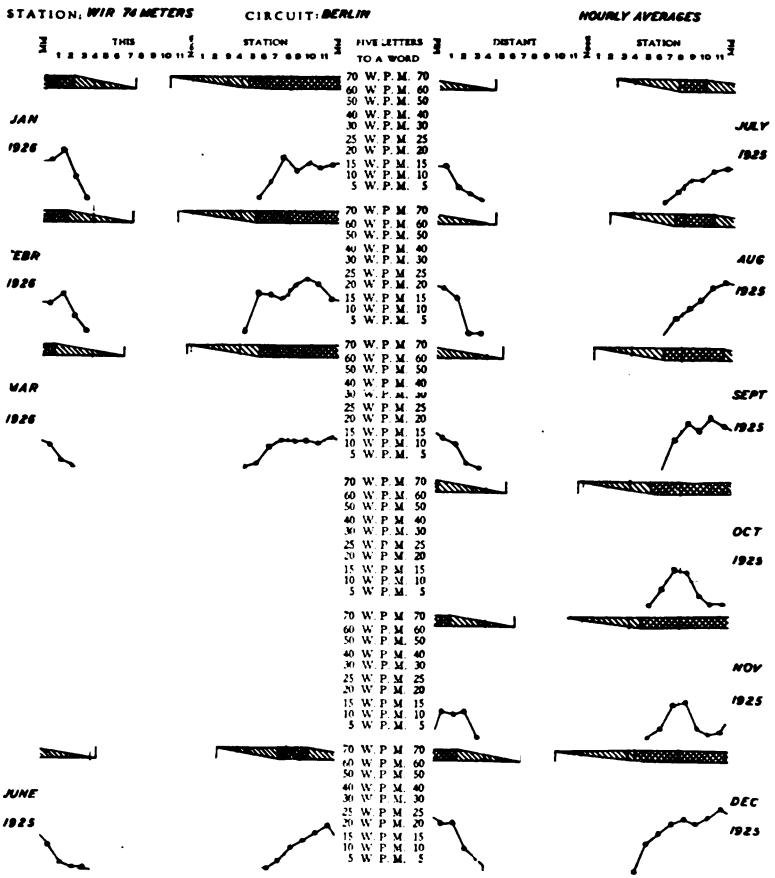

Figure 10—Commercial reception of New Brunswick—WIR, at Berlin, June, 1925 to March, 1926.

cial traffic was WGH, on 103 meters, which went into service in September 1924. As shown in Fig. 8 it was found fairly effective for very slow transmission to Buenos Aires between the hours of 8 P. M. and 6 A. M. This period during October advanced with earlier sunset here, becoming 6 P. M. to 5:30 A. M. However the coming of winter at New York

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in November and December meant the approach of summer at Buenos Aires, with its heavy static season. Signals during November became less and less reliable until during December the usefulness of this transmitter to Buenos Aires had passed and thereafter it was hardly audible and never readable.

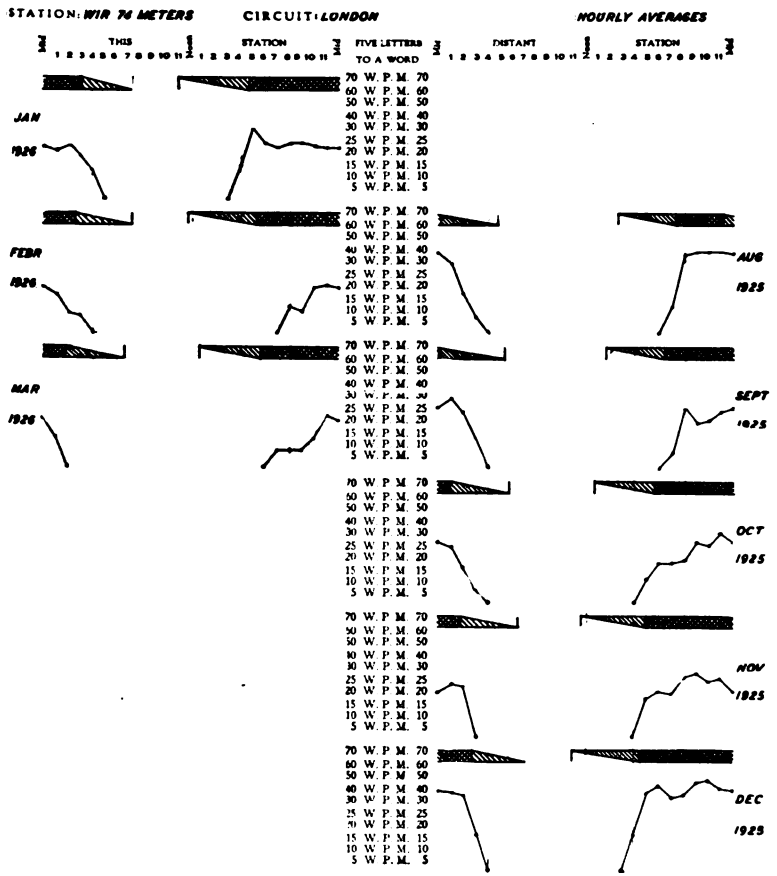

Figure 11—Commercial reception of New Brunskick—WIR, 74 meters, at London, August, 1925 to March, 1926.

In the meantime Berlin had become extremely interested in the use of short waves and was receiving traffic from WGH wherever that station could not be used to Buenos Aires. Fig. 9 shows the record of improved reception at Berlin during the winter months of 1924-5.

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However, the approach of spring brought a sharp decline in reception speed and a shortening of the readable hours. By the end of May, Berlin was unable to accomplish much commercially and in June 1925, this pioneer transmitter, having served its purpose, was dismantled.

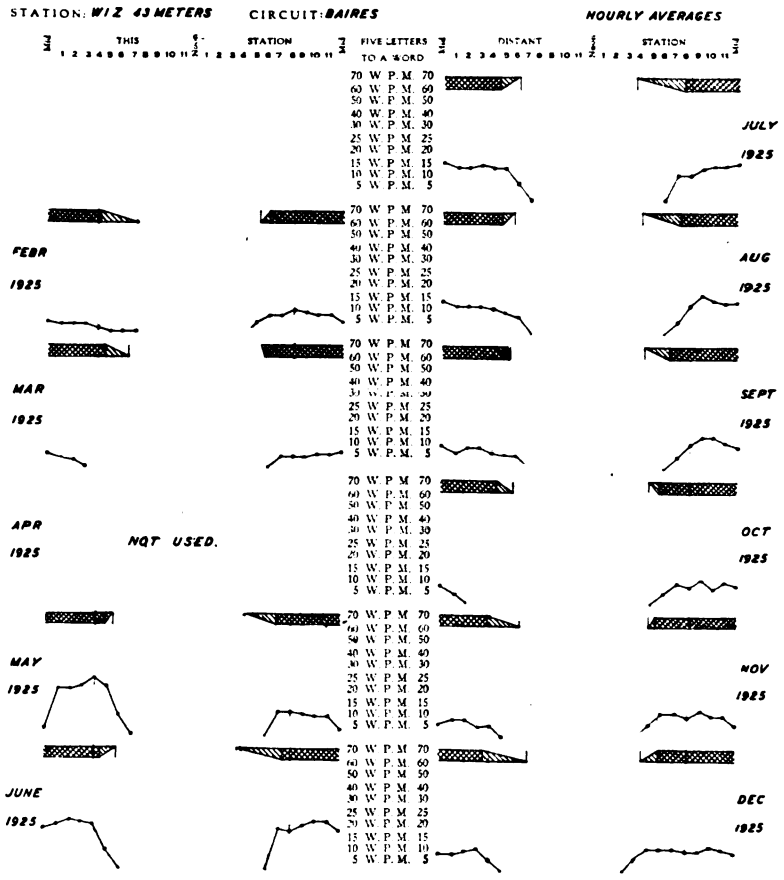

Figure 12—Commercial reception of New Brunswick—WIZ, 43 meters, at Buenos Aires, February to December, 1925.

Next in order of wavelength came WIR, 74 meters, located at New Brunswick, New Jersey and beginning operation in June, 1925. This transmitter was not used to Buenos Aires, because in May, 1925, a 43-meter transmitter, WIZ, had gone into service and was proving far superior for

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South American service to the 103-and 74-meter wavelengths.

WIR did however operate successfully to Berlin and to London during almost a year. In hours of usefulness and speed of operation to Berlin, it proved to be comparable to WGH, perhaps becoming readable some one-half to one hour earlier in the day and being slightly less subject to daily and seasonal static conditions.

London did some very good work in reception of this transmitter, copying at high speed for hours at a time and receiving over a somewhat longer period each day than did Berlin. Shorter distance and somewhat longer hours of darkness probably account for this difference. Results of WIR transmission to Berlin and to London are given in Figures 10 and 11.

This station, although very useful in the movement of traffic to Europe, was discontinued after March, 1926 in order to furnish apparatus for shorter wave-length transmitters for South American service.

Station WIZ, 43 meters, at New Brunswick, went into operation in February, 1925. It has become perhaps the best known short wave transmitter in the world, being used by amateurs, commercial administrations, and military services, as a standard of comparison for reliability and wavelength. It is heard over the entire earth at some hours of each day.

During 1925 it was subject to innumerable changes in design. Traffic data for that year are therefore subject to much question and explanation. Results for 1926 are more reliable, the design and operation of the transmitter having become more stable and well standardized. This is the only transmitter for which two years data is available and the natural tendency is to compare 1925 with 1926. Much of the discrepancy between the two years is thus explained.

The 43-meter wavelength proved much less subject to seasonal effects for use to Buenos Aires than had the previous longer waves, and therefore was more reliable. Speed of transmission to Buenos Aires is normally quite low. To Berlin very good high-speed operation is obtained and at least a fair commercial speed is to be depended upon practically every night of the entire year.

Periods of weak signals and strong signals of several

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days or weeks duration which affect reception heavily are reported by Berlin. These periods have been studied in relation to sun spots, magnetic disturbances, barometric pressures, and other natural phenomena, but no definite

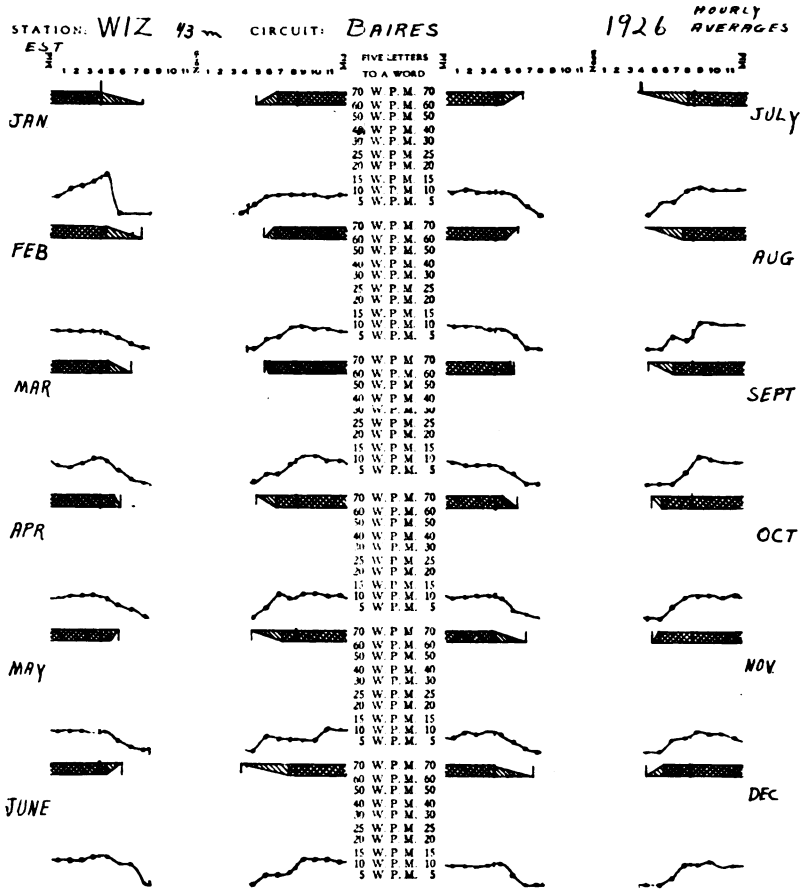

Figure 13—Commercial reception of New Brunswick—WIZ, 43 meters, at Buenos Aires for year 1926.

connection is yet established with anything. A more or less regular cycle of about 35 days is becoming noticeable and is being checked up at present.

Hours of readability during 1925 and 1926 to Buenos Aires and Berlin are shown in Figs. 12, 13, 14 and 15. Although 43 meters is a "darkness" wavelength for all transoceanic operation, the readable period does not follow

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seasonal variations of sunrise and sunset as closely as might be expected.

Stations WQN, 51.5 meters, and WQO, 35 meters, both located at Rocky Point, New York, were operated commercially for several months during the fall of 1925. The results

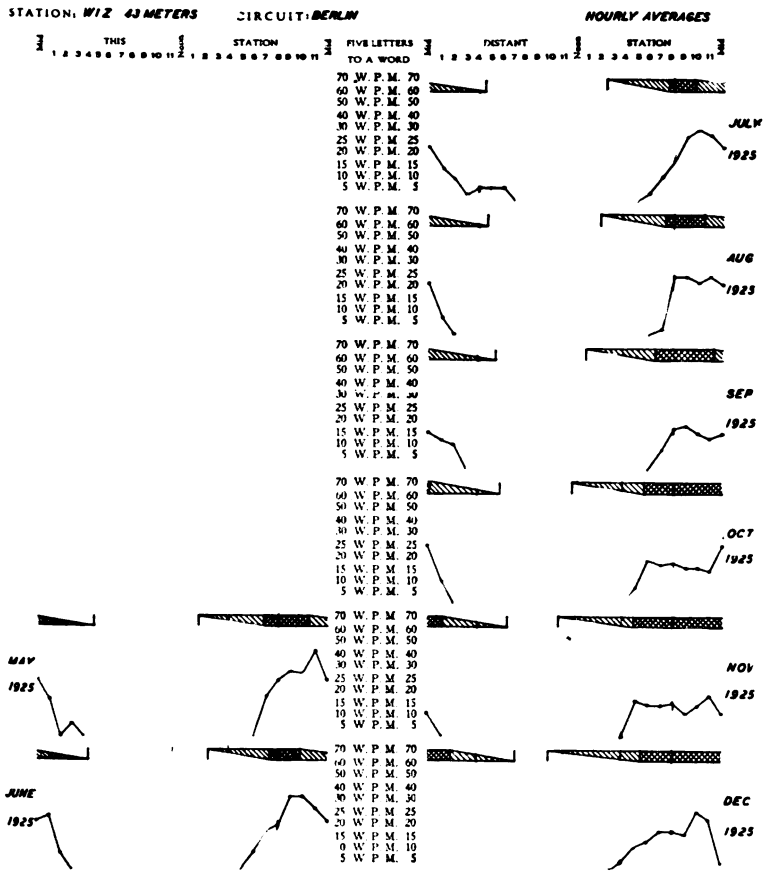

Figure 14—Commercial reception of New Brunswick—WIZ, 43 meters, at Berlin, May to December, 1925.

obtained were so nearly identical with those on 43 meters that operation of the Rocky Point transmitters was discontinued and efforts concentrated on an attempt to utilize much shorter wavelengths which were expected to provide daylight operation to supplement the darkness operation on 43 and 74 meters.

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One of the most important uses made of WQO was a run of three weeks using a wavelength of 44 meters and output of about 12 kw. for comparison with WIZ on 43 meters and output of about 6 kw. It was determined without doubt that the traffic capacity of the two transmitters,

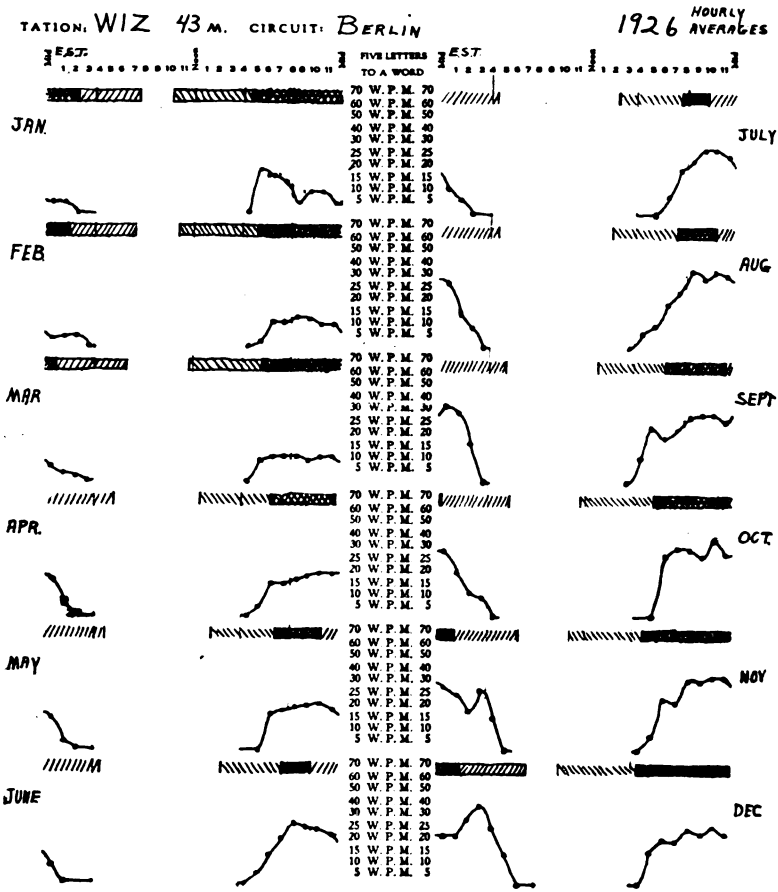

Figure 15—Commercial Reception at New Brunswick—WIZ, 43 meters, at Berlin, for year of 1926.

wavelengths being nearly alike, was in direct proportion to their power. This disposed of a theory which had gained much credence that in short wave working the effectiveness of a transmitter was determined entirely by its wavelength and that power increase would be of negligible effect.

2XS, 14.9 meters, at Rocky Point, went into service in

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January, 1926 and during the entire year was operated successfully and very regularly to Buenos Aires and Rio de Janeiro during daylight. Fig. 16 gives readability records of 2XS, 14.9 meters, to Buenos Aires. The beginning of readability each day coincides quite sharply with sunrise at New

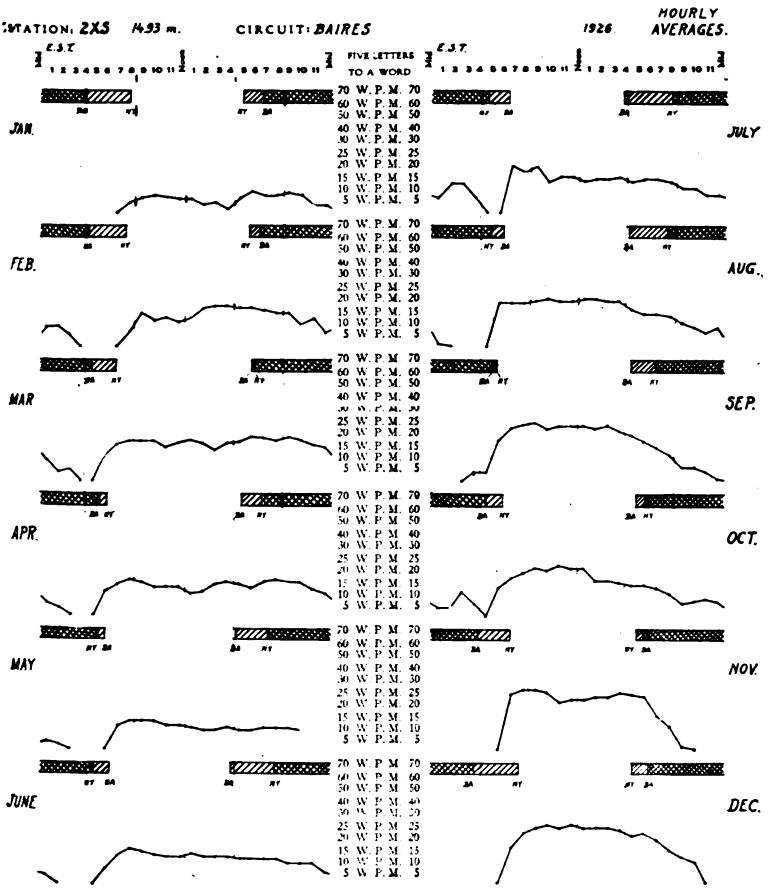

Figure 16—Commercial reception of Rocky Point—2XS, 14.93 meters, at Buenos Aires, for year of 1926.

York but continues after sunset at both New York and Buenos Aires and varies greatly in its "fading" time according to season of the year. Hours of readability at Rio de Janeiro very closely agree with those shown for Buenos Aires.

European reception of 14.9 meters is confined to a very

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brief period each day. Due to constant operation of the 2XS transmitter for South American service no regular use of it has been made to any European point and but little operating data secured on this wavelength for such use.

Other transmissions have been conducted using trans-

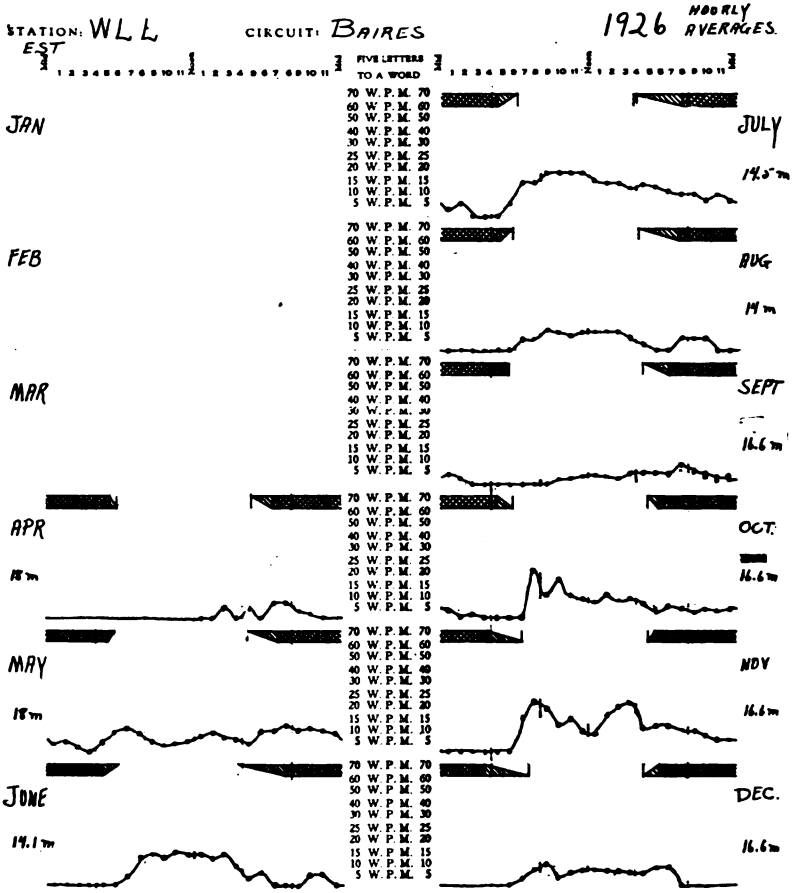

Figure 17—Commercial reception of Rocky Point—WLL, 18 to 14.1 meters, at Buenos Aires for year 1926.

mitter WLL at Rocky Point, at first on 18 meters during April and May 1926. Readability to Buenos Aires covered practically the full 24 hours but signals were weaker and less reliable than 2XS. This may be attributed to the more experimental nature of the transmitter installation rather

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than to difference in wavelength. Record of WLL operation to Buenos Aires is given in Fig. 17.

Any account of short wave operation to date must include the 22-meter station WIK at Rocky Point although its operating data include only two months, November and December 1926.

This wavelength is peculiar in being useful to South America as a "darkness" transmitter and to Europe during daylight so that it is in service practically 24 hours per day. Buenos Aires reads 22 meters from 5 P. M. to 8 A. M. E. S. T. and Berlin receives it between 6 A. M. and 4 P. M. E. S. T. Results to Berlin are also marked by reception on 22 meters during about four hours daily of "double signals" which seem to be caused by reception of the normal signal over the direct path New York to Berlin and also reception of the same signals some ⅛ to 1/10 second later arriving from the opposite direction around the earth. These phenomena are also observed on shorter waves. This subject has been given serious study by Mr. Erich Quaeck of Berlin, whose conclusions are published in an article appearing in the December, 1926 issue of "Jahrbuch der Drahtlosen Telegraphie und Telephonie". The difference in time of signal travel, is such that useful reception of traffic was impossible because of the "double signal" filling in the spaces between dots of the normal signal. A directional receiving system has been used to overcome this difficulty.

Short wave operations on the Pacific have in general confirmed the data and results obtained in trans-Atlantic operation. Conditions there have been such that regular and systematic use has not been carried out to the same extent as on the east coast and therefore no data can be submitted until further operation has afforded more opportunity to secure reliable and comprehensive information.

No mention is made in this paper of theories of short wave transmission and no attempt made to state a theory derived from the results obtained, nor to explain such results in terms of already propounded theories. The intention has been merely to record what has been done in this field commercially in the belief that such information will be found of benefit to those who are engaged in either theoretical or practical consideration of short wave developments.

In Section III, there is recorded a number of the prob-

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lems encountered in the technical development of short wave transmitters.

III—Technical Development of Transmitters

The introduction of short waves for long distance communication has brought into being a technique widely different from that used in connection with the older long wave systems. The radio frequency currents are no longer generated directly in a dynamo-electric machine like the Alexanderson alternator but are produced entirely by means of vacuum tubes.

In contrast to the antennas for long waves, four hundred feet high and one and one-half miles long, we have for short waves antenna structures so small and insignificant in physical proportions that a casual visitor to a transoceanic station would not think of taking them seriously if he were not familiar with their effectiveness.

The revolutionary change in methods has presented a multitude of problems and difficulties requiring solution by the engineers engaged in the development. Most of these difficulties have had to do with the production of equipment capable of producing and handling the extremely high frequency energy in a manner suitable for communication purposes.

The investigation of short waves for long distance transmission was delayed for so many years because of the lack of equipment capable of producing sufficient power at very high frequencies. Utilization of short waves on a practical manner has depended almost entirely upon the development of vacuum tubes.

The technical development described has resulted from the close cooperation between the engineers of the RCA and those of the General Electric and Westinghouse Companies, particularly those responsible for the development of tubes and transmitters.

When powerful tubes capable of producing very high frequency currents were made available the development of short wave communication became so rapid that the engineers engaged in developing the tubes have been under constant pressure to make their product more and more adaptable to short wave operation. From the start the de-

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velopment of tubes has set the limit for the development of short wave transmitter equipment and the life of the tubes in service has proven to be one of the greatest single factors in the economics of the new system.

It is believed that most of the difficulties encountered in using the commercial tubes at very high frequencies have been due to the high radio frequency currents which flow through the connections to the tube elements and to the generally less efficient or more unstable circuits used in connection with the tubes.

The high currents through the tube connections are due to the dielectric capacities between the elements which become quite low reactances at high frequencies. These currents increase the temperature of the elements, particularly the grid, resulting in a tendency for the tube to become unstable due to electron emission from the grid and the higher temperatures combined with the severe conditions which the tube imposes upon itself due to instability, tend to evolve gas from the elements and destroy the usefulness of the tube.

The circuits used subject the tubes to increasingly severe conditions as the frequency is raised due to the necessity for increasingly accurate neutralization of the coupling between input and output circuits through the dielectric capacity of the tubes. The probable error from exact balances of this coupling increases in proportion to the frequency and increases the probability of the tube being subjected to too little or too great excitation or oscillation under improper conditions which tend to destroy it. Variation in temperature of tubes and circuits is alone sufficient to introduce serious difficulty in maintaining exact neutralization of the tube capacity in the ordinary radio frequency amplifier when used at wave lengths as short as fourteen or fifteen meters.

At first trouble was experienced due to the high temperature of the tube connections and to dielectric losses causing softening of the glass resulting in glass punctures but this trouble has been very greatly reduced by improvements in design.

It is also to be noted that the tube capacities set a limit to the capacity of the oscillating circuits used in the input and output of the tubes, and therefore determine the watt-

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less current circulating in these circuits. For very short waves the amount of this circulating current is of necessity greater than is desired and results in excessive circuit losses. In fact, if waves very much shorter than fourteen meters are used with the present tubes, most of the normal output of the tubes is lost in their associated circuits and nothing left over for the antenna.

It is significant that within about two years, due to the development of both tubes and circuits, the average life of tubes has increased in the ratio of about three-to-one.

Although the development of vacuum tubes has undoubtedly been the most important and most difficult problem in the commercial application of short waves it has been by no means the only one.

The maintenance of a constant frequency became a serious problem requiring solution. To obtain maximum usefulness the transmitters used should be capable of reception within a frequency band of about the same number of cycles as when long waves are used. To stay within this band requires that the percentage of frequency variation in both the transmitter and the receiver decreases in proportion to the wavelength. Thus for communication on a wavelength of 15 meters it is necessary to maintain a constancy about one thousand times greater than was required for long waves.

This problem has been solved by the adoption of the piezo-electric quartz crystal oscillator which was first used as a master oscillator in transmitter 2XAO at Belfast, Maine and which had proven quite satisfactory as a source of constant frequency. As a result of the use of the crystals, together with various refinements in their operation, the frequency of the transmitters and receivers can be held so constant that, the heterodyne note stays sufficiently fixed for hours, or even days without any adjustment to the equipment.

A further advantage resulting from the use of the quartz crystal master oscillator is the comparative simplicity and consequent low cost of the transmitter. This results from the fact that the output frequency is made independent of the effect of variation and ripple in the power supply, and alternating current may be used for lighting the filaments of the tubes. This brings about a very great saving in the

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amount and cost of the equipment used to make up the transmitter.

Next in importance to maintaining the frequency constant was the problem of amplifying the very small output from a crystal oscillator to the power required in the antenna in a manner which would give freedom from feed-back from the higher to the lower power stages and the resulting tendency for undesired oscillations. It has been

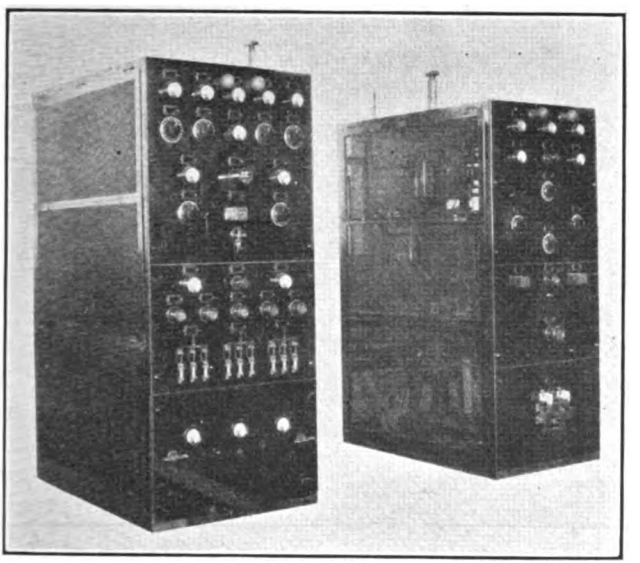

Figure 18—Short wave transmitter for Bolinas—14.1 meters, showing operating condition.

previously found difficult to operate a chain of amplifiers even on relatively long waves without feed-back troubles, and the difficulties increase with the frequency. Shielding can be used to reduce such troubles but can not be relied upon to eliminate them entirely at very high frequencies. Also it was necessary that the effect of coupling between input and output circuits of each stage of amplifier be small and that there be no critical adjustments to get this result. In addition it may be noted that it is not practical to make and operate crystals at wavelengths short enough to be used to control the transmission directly, it being necessary to use a harmonic of the crystal.

These requirements were met by using each of the in-

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termediate stages as combination amplifiers and frequency multipliers. This was done by tuning the input circuit of the tubes to some particular frequency and the output circuit for the second or third harmonic of the input frequency. Due to the great difference in tuning between the input and output circuits the coupling between them can cause only a very small amount of feed-back and the neutralization of the tube capacity is no longer critical. Also the feed-back from the

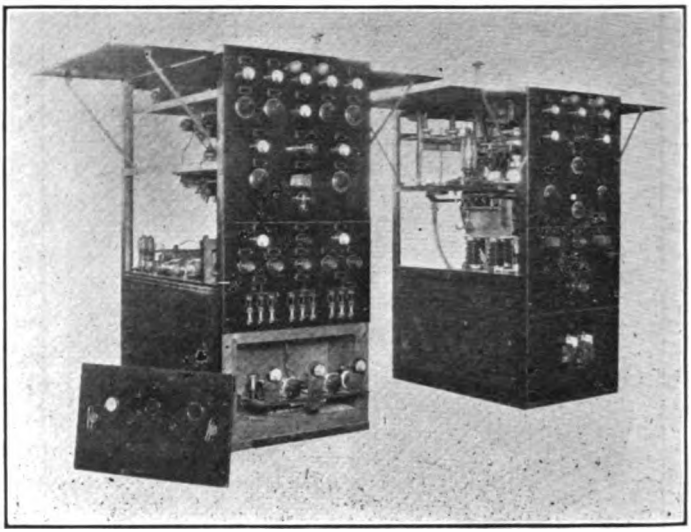

Figure 19—Short wave transmitter—14.1 meters, shields open showing accessability of parts.

higher to the lower power stages cannot set up oscillations because of the very great difference in operating frequency.

Incidentally this system allows the greater part of the tubes in the system to operate at relatively low frequencies so that they are easier to handle and are not subjected to such heavy capacity currents.

In using a tube as a frequency multiplier, advantage is taken of the fact that when it is used as a high frequency oscillator or amplifer the anode current flows for only a small part of a half cycle and can be resolved entirely into harmonic components of the input frequency by tuning the output circuit for the second or third harmonic. In doing so there is only a little loss in efficiency and, due to the greater stability of the circuits, it is believed possible to be confident

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of obtaining in a practical manner, a greater output than can be relied upon if the tube is used as an ordinary amplifier at very high frequencies.

A third problem of major importance which requires solution before short waves can be used to full advantage for commercial communication is the reduction or elimination of fading. Up to the present time this problem has not been completely solved but some work has been done with promising results.

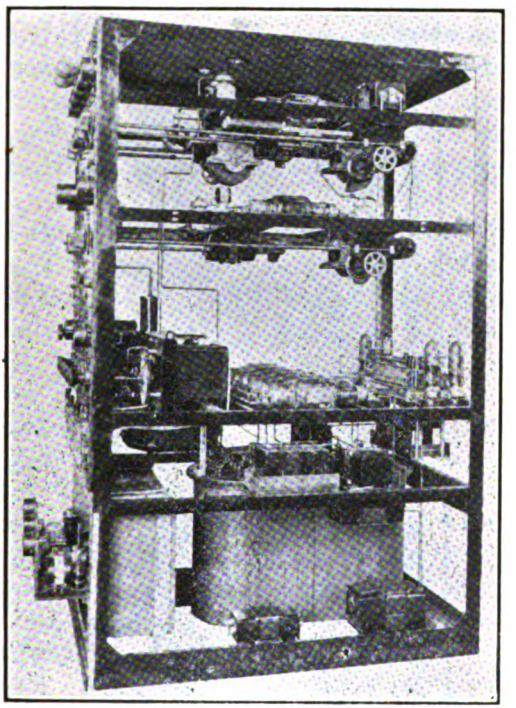

Figure 20—Short wave transmitter—14.1 meters, master oscillator and intermediate amplifier unit, side view.

Very early in the development of short waves it was observed that the fading did not occur simultaneously on two different frequencies transmitted from the same point simultaneously even though the difference in frequency might be very small. This immediately led to the conclusion that fading could be very greatly reduced by transmitting a band of frequencies instead of confining all the energy to a single frequency. This may be done by modulating the transmitted waves or by continuously varying their frequency. Both methods have been tried and were found to

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result in some improvement, although in no case was the fading eliminated. Since most of the receiving stations were required to transmit the signals over wire lines with some noise on them they preferred a single steady frequency which would give a pure note at the receiver easily distinguishable from line noises. The present transmitter models therefore do not have any provision for varying the frequency over a narrow band while keeping the average frequency fixed.

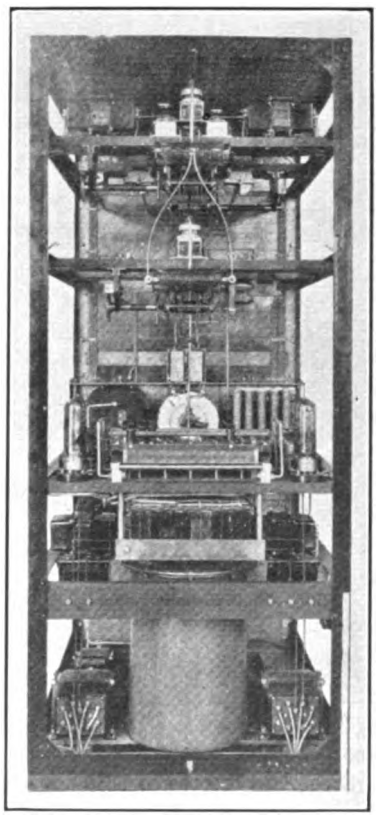

Figure 21—Short wave transmitter—14.1 meters, master oscillator and intermediate amplifier unit, rear view.

If the time comes when conditions at the receiving stations permit taking the variable frequency signals it will no doubt be found an advantage in some cases to do so.

Work which has been done in connection with reception promises much more important results than anything which can be done at the transmitter but this work is still under way and will be described at a later date.

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The use of short waves has made it possible to construct directive antenna systems at a cost which is not prohibitive and the further development of such systems seems certain to bring about very great improvements in the effectiveness in the short wave transmitters now in commercial service. of short waves. A program of installation and development of directive antennas is under way; but a discussion of this subject would be too lengthy to be included in this paper.

Figures 18 to 23 inclusive are illustrative of a practical form in which the various developments have been embodied The particular equipment illustrated operates on a wavelength of 14.1 meters and was constructed for installation at Bolinas, California. Only the radio frequency units of the transmitter are shown. In addition to these units each transmitter includes a high voltage rectifier of conventional design.

Figure 15 shows a general view of the transmitter units totally enclosed as they would be used in service. Fig. 19 illustrates the accessibility of all parts. Fig. 20 gives a view of the assembly of the master oscillator and intermediate amplifier unit, while Fig. 21 gives a rear view of the last stage or power amplifier unit.

The master oscillator and intermediate amplifier unit forms a complete crystal controlled short wave transmitter having an output of 300 to 400 watts. It contains all the rectifiers necessary to supply anode and grid bias voltages. To make it ready for service requires the connection of only three leads to deliver 220 volts, three phase, 60 cycles, two leads for the telegraph control and two leads for the radio frequency output.

The crystal oscillator and first combination frequency multiplier and amplifier are in the metal box at the lower front of the unit where they are shielded from the other stages and where their temperatures may be automatically controlled when desired. RCA type UX-210 tubes are used in these two stages.

Two additional stages of combination frequency multiplier and amplifier using RCA UV-204A tubes are located in the upper part of the unit. There are two rectifiers, one using four UV-217A tubes and another using six UV-1651 tubes as a three-phase full-wave rectifier.

The keying is accomplished through the agency of a high-

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speed polarized relay which controls the grid bias potential on the UV-204A tube which is used as the second stage intermediate amplifier.

The power amplifier unit is designed to use two water-cooled tubes in the ordinary push-pull amplifier circuit. No

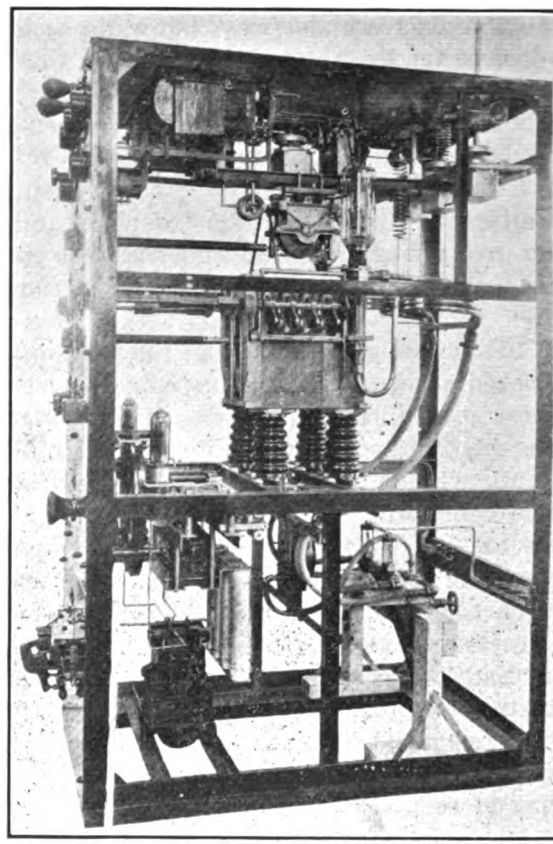

Figure 22—Short wave transmitter—14.1 meters, power amplifier unit, side view.

change in frequency is made in this stage although there is some reason to believe it would be more satisfactory if it were used as a frequency multiplier.

The unit contains its own rectifier for supplying grid bias potential to the water-cooled tubes consisting of two UV-217A tubes used for single-phase full-wave rectification.

It may be of interest to note that this type of amplifier

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has been designed to have the cooling water to the tubes fed through the conductor of the output circuit coil with rubber hose connections near the center of the coil. This very greatly reduces the radio frequency losses in the water

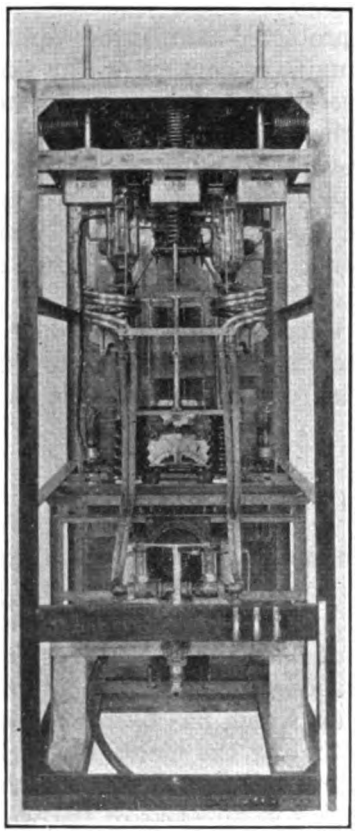

Figure 23—Short wave transmitter—14.1 meters, power amplifier unit, rear view.

connection because the center of the coil is substantially at zero potential radio frequency.

It has been definitely determined that the formation of scale on the anodes of water-cooled tubes, a condition so frequently encountered in practice, is due to overheating of the anode accompanied by the formation of steam in the water jacket. Boiling has been eliminated by providing an adequate flow of cooling water with the result that the anodes

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may be expected to remain clean and bright throughout the life of the tubes.

Protection is provided against loss of cooling water pressure and flow by means of a contact-making pressure gauge and by two contact-making flow devices placed in the water outlet from each tube.

The tubes are protected against too rapid starting of the filaments by an auxiliary contact on the filament rheostat which makes it necessary for the operator to insert all the resistance before the voltage can be applied. A second auxiliary contact prevents application of the anode voltage unless the resistance is reduced again to approximately the normal value and the switching is so arranged that the anode voltage can not be applied until after the grid bias is on.

Very great care has been used to reduce all stray capacities and the inductance of connections to a minimum. This feature becomes increasingly important as the frequency is increased.

IV—Acknowledgment

Short wave communication is unique in that it is world-wide in its scope. A comprehensive mental picture of progress in this field of human activity involves the receipt and accumulation of data from all corners of the earth. It is obviously impractical to acknowledge individually, the splendid cooperation which has been extended by the various departments, and by the manufacturing companies of the Radio Corporation of America, or by its remotely located traffic correspondents in all parts of the world; but rather to express to all, appreciation for data and suggestions submitted from time to time. It is felt that communication by short waves will become an increasingly effective means for welding all the peoples of the earth into bonds of mutual understanding.

Synopsis: The development of short wave communication by the Radio Corporation of America is outlined. A summary of short wave installations, with call letters, wavelengths and services to which each installation is assigned, is submitted.

Traffic charts showing the diurnal and seasonal char-

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acteristic of various wavelengths over typical circuits are shown.

An outline of the technical problems inherent to the development of tubes and transmitter circuits is discussed. Methods are described for obtaining proper operation of tubes and transmitters at these very short wave lengths.

The paper is illustrated with typical pictures and charts showing transmitter development and traffic performance.

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SOME PRACTICAL ASPECTS OF SHORT-WAVE OPERATION AT HIGH POWER*

By
H. E. HALLBORG

I—PROPAGATION CHARACTERISTICS

The art of short wave transmission is so closely linked with the phenomena of propagation at these extremely high radio frequencies, that any contribution to our present all too meagre knowledge of the subject of short wave phenomena will doubtlessly be acceptable. The wave bands covered by this paper are those included within the 10-to 100-meter zone, or those having a frequency range of 3,000,000 to 30,000,000 cycles.

Early experiments with these frequencies produced results so erratic and contradictory that no symptoms of dependable laws of progapation were in evidence. Continued experimentation by many workers in this field however, indicates that short wave propagation is a wholly orderly process. It appears that nature again has refuted the challenge of being slipshod in any of its methods.

Evidence is increasingly cumulative that energy radiation at these frequencies is divisible into two wave groups and an intermediate zone as follows:

(a) A highly attenuated ground wave which follows closely the established law of attenuation at long waves.

(b) A slightly attenuated sky wave projected from the antenna at an elevation which insures that its transmission path is unobstructed by earthbound obstacles, traversing almost entirely the rarefied regions of the upper atmosphere.

(c) Intermediate between these two zones is one in which the signal becomes very weak, or entirely non-existent. This is the skip distance zone.

The nature of the propagation characteristic of short waves for distances not in excess of 3,000 miles have been

*Received by the Editor April 4, 1927.

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studied at the Radio Experimental Station of the General Electric Company at South Schenectady, New York. Figure 1 is a plot of the day and night signal audibilities as observed at South Schenectady for wave lengths of 26.4, 32.79, 50.2, 65.5 and 109 meters. A radical difference in propagation efficiency is indicated for both day and night observations. Skipping was observed on the shorter waves 26.4 and

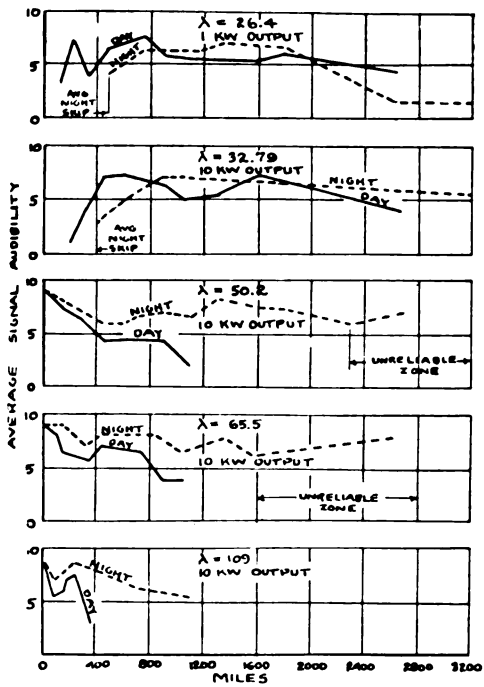

Figure 1—Day and night signals at various wavelengths.

SIGNAL CODE

0—No signal.	3—Readable.	6—Fair.	9—Strong.
1—Just audible.	4—Faint.	7—Good.	10—Very strong.
2—Unreadable.	5—Weak.	8—Loud.	

32.79 only. The skip at night exceeds the day skip by a ratio of nearly two-to-one. Unfortunately these tests did not extend below 26.4 meters into the region where the skip distance and diurnal effects are most marked. It has been observed by many investigators that increase of wave frequency results in increase of skip distance, and decrease of ground wave range.

The fact that a skip distance actually exists indicates that the energy is radiated at a considerable elevation. The

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further fact of long skip with shorter wavelengths indicates a relation between elevation and frequency. It is conceivable that the angle of projection may become so high that the wave front does not again come down to earth. This appears to be a plausible explanation for the ineffectiveness of frequencies greatly in excess of 30,000 kc. (10 meters)

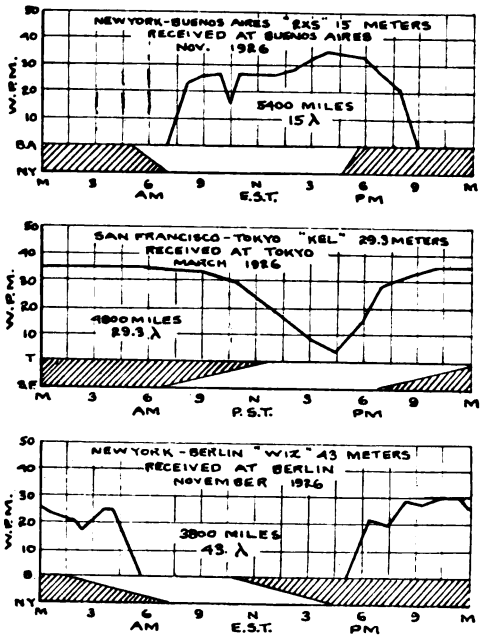

Figure 2—Diurnal signal variations on three typical wavelengths and traffic channels.

day or night. Even at frequencies as low as 15,000 kc. (20 meters) the atmosphere absorbtion at night is normally not sufficient to refract the wave front sufficiently to bring it down to earth. It will be noted that the physical mechanism of propagation may be accounted for without reference to an upper reflecting layer, but rather by the combined effects of wave frequency and the relative ionization of the atmosphere.

It is to be expected, if the above hypothesis is tenable, that light and darkness with their equivalent high and low ionization values will have a marked effect upon the strength of short wave signals.

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This diurnal effect is well illustrated by the characteristic hour by hour signal intensities observed on 3 typical long distance circuits of the Radio Corporation world network as illustrated in Figure 2. It will be noted that the 15-meter signal is most effective during the daytime or during the period of high ionization. The overlap of 3 hours between 6 and 9 P. M. may possibly be explained by residual

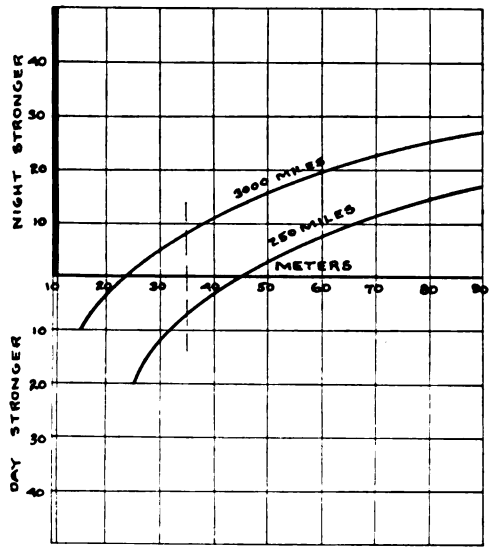

Figure 3—Relation between distance and day waves vs. night waves at 250 and 3000 miles.

ionization of the atmosphere since it was summer at Buenos Aires when the data were obtained.

The 43-meter signal is seen to be effective only during the hours of low ionization or when the transmitter is completely in darkness.

Intermediate between these extremes must be a frequency which is less susceptible to relative atmospheric ionization. This may account for a characteristic somewhat similar to that of the San Francisco-Tokyo result shown in the middle graph of Figure 2.

It will be apparent from the data already presented that certain wavelengths are most effective at night, whereas others are only effective in the daytime. This diurnal wave characteristic is also to a certain extent a function of dis-

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tance. For instance, reference to Figure 3, shows that at a distance of 250 miles waves ranging from 45 meters and above are stronger night than day signals. At three thousand miles, wavelengths above 24 meters are night signals and below 24 meters have stronger day than night signals. The 35 meter wave may develop a day signal at 250 miles and a night signal at 3000 miles. This condition is indicated

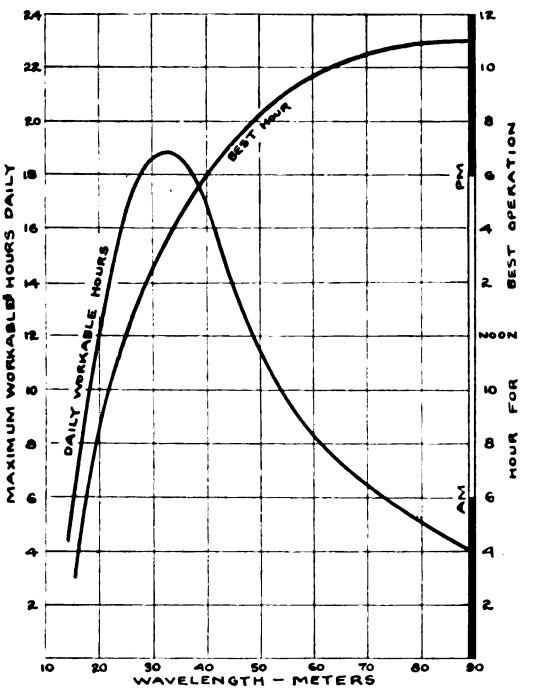

Figure 4—Diurnal characteristics for short wave transatlantic operation.

by the dotted vertical line of the figure. While the values submitted are not claimed to be exhaustive and invariable, they do indicate a definite trend which is being checked up by daily experience.

Many data have been accumulated by the Radio Corporation of America relative to the number of workable hours and the best hour for trans-Atlantic operation as a function of wavelength. These data are based upon an input of 10 kilowatts into a single wire vertical radiator. The results are plotted in Figure 4. It will be observed that a

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33-meter wave gives the maximum number of hours of operation for trans-Atlantic working. It is also indicated that the day time band for this distance lies between 17 and 40 meters. It is also of interest to observe that with a non-directional type of antenna and 10 kilowatts antenna input at least two waves will be required to insure 24 hours of trans-Atlantic operation.

In further evidence of the fact that relative atmospheric absorption may account for the day and night effects above

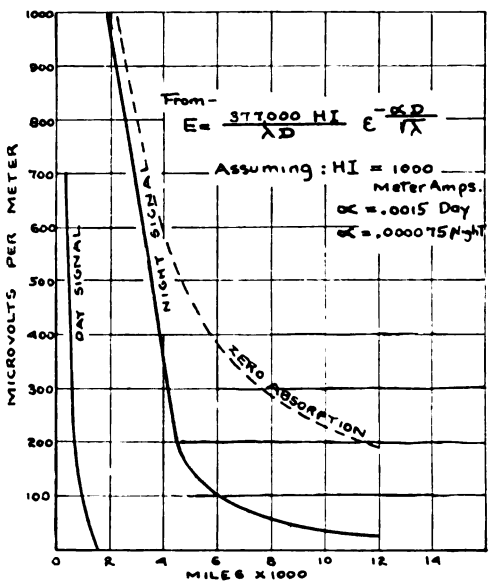

Figure 5—Calculated signal strength—100-meter wave over salt water.

noted, signal strength versus distance characteristics have been calculated and plotted under conditions as shown in Figure 5. The daytime absorption coefficient value .0015 is that normally assumed at long waves over salt water. The night time coefficient .000075 is chosen as a result of observed experimental data on the 100-meter wavelength. The dotted curve indicates signal strength with zero atmospheric absorption assuming the signal to obey the inverse law only. Of particular interest is the "hockey stick" shape of the night signal characteristic. This indicates one possible explanation of the distance effectiveness of short waves

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in view of the remarkably good signal-to-static ratio which always obtains in the short wave band.

There is every indication that the Austin-Cohen transmission formula will be applicable to the short wave band when experience has indicated the normal values of the attenuation constant to be applied to each band of 1000 kc. The new values of this constant will probably lie between .000075 and .0005.

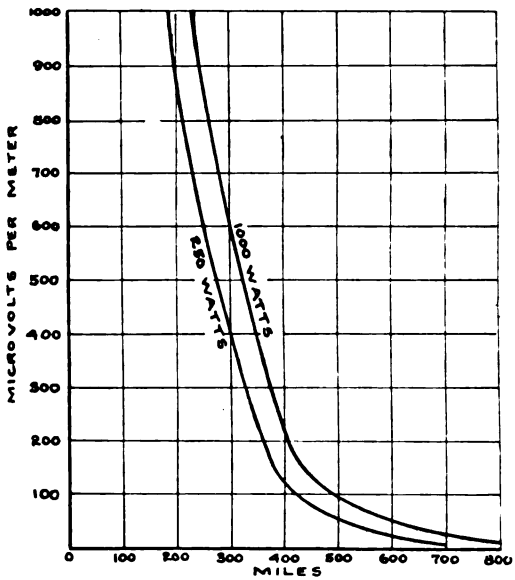

Figure 6—Comparative strength of signals calculated for 250 and 1000-watt antenna inputs at 29.5 meters $\alpha = 0.00075$.

Figure 6 is of interest in showing the relatively small calculated increase in signaling distance at 29.5 meters when the antenna power is increased, in the ratio of 4 to 1 namely from 250 to 1000 watts. This calculated result is well in agreement with practical experience.

II—Frequency Stabilization

One of the problems of first importance in short wave operation is that of holding constant frequency. In order better to realize the significance of this problem it is sufficient to contemplate that the range of average audibility is

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usually taken as 20,000 cycles. If we assume that transmission is on a wavelength of 15 meters, or 20,000,000 cycles, then a frequency change of the order of one percent represents a cyclic change of 200,000 cycles. This one percent change is ten times the entire audible range. A change equal to the audible range is equal to a shift of frequency at 15 meters of only 1/10 percent. Good practice requires that the frequency shift shall be not in excess of 100 cycles. This means that at 15 meters a frequency change of

$$\frac{100}{20{,}000{,}000} = .0005 \text{ per cent is the maximum permissible.}$$

Of the various factors that contribute to stabilization of frequency the following may be enumerated:

(a) The filament emission must be kept at a constant value. A well regulated d-c. source is better than a-c. although obviously less flexible. Tests conducted on the West Coast by the writer in which the filaments were transferred from an a-c. to a d-c. source, all other conditions the same, proved that frequency shifts on a 29.3-meter transmitter could be introduced merely by the slight emission variations incident to the use of an a-c. source instead of d-c.

(b) The plate voltage supply must be well smoothed and the load regulation must be good. A ripple value of 1% or less, with a regulation not in excess of 10% may be considered good practice. In extreme cases it may be necessary to provide an auxiliary load which acts to keep the rectifier load constant under all conditions.

Thermal expansion in the coil or condenser units may change the circuit constants. This trouble may be eliminated by conservative design which results in lowering the unit stresses.

(d) Finally the frequency may be kept constant by piezo-electric crystal control. A quartz crystal has the property of oscillation of a fixed frequency determined by thickness. This oscillation frequency is maintained during the life of the crystal providing its thickness is not altered by thermal expansion. To insure constant temperature the crystal may be mounted in a separate container the temperature of which is controlled by a thermostat.

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III—Comparison of Crystal Control Circuits

All circuits which utilize crystal control are similar insofar as they attempt to impress upon the antenna energy the frequency stability of the crystal. However, since the energy available at the crystal is a few watts at most, and that to be controlled several thousand watts, it is obvious that there are many pitfalls in the train of circuits from crystal to antenna. For this reason not all crystals succeed in justifying their prime reason for existence.

Crystal control circuits may be divided into three classes as follows:

(1) Crystal amplifier
(2) Harmonic selector
(3) Frequency doubler.

The crystal amplifier, Class (1), is distinguished from the others, in that the frequency of the crystal, and that at the antenna are the same. This means that the energy of the crystal must be applied to the antenna through a cascade system of straight amplification. There is consequently a practical limit to which this method can be applied which is determined by crystal thickness and life. Crystals are being ground for frequencies as high as 10,000 kc., or 30 meters, but for long life the use of a crystal ground to 3,000 kc., or 100 meters ,or above is preferable.

The harmonic selector circuit, Class (2), utilizes a thick crystal oscillating at 100 meters or above. This 100 meter oscillation is then fed to the grid of a selector circuit, the value of grid bias of which is adjusted so that the third, fifth or seventh harmonic as desired may be selected. The final frequency of operation is obtained in the selector stage, which is followed by a system of straight amplification. The energy available in the selector stage is roughly inversely proportional to the harmonic selected.

The Frequency Doubler Circuit Class (3) also utilizes a thick crystal oscillating at 100 meters or more. This 100-meter oscillation is then fed to the grid of a second stage the bias of which is adjusted to maximum amplification of the 2nd harmonic, or double frequency value. The second stage would thus oscillate at 50 meters assuming a 100-meter crystal. The second stage may then be similarly fed to a third stage adjusted to select the 2nd harmonic of the sec-

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ond stage namely, 25 meters, and so on. This system has the advantage of reducing direct coupling effects between stages and the disadvantage of less perfect control of regeneration.

Perfect control of the antenna energy by crystal requires that there shall be no parasitic oscillations in any of the amplifier stages. This requires that each stage shall have sufficient energy to drive the next succeeding stage without resort to regeneration.

IV—Features of Design and Adjustment of 20-Kilowatt Amplifier

It is obviously not possible in a brief summary of this nature to do more than indicate a few of the most essential features of the design or adjustment of a high power short wave power amplifier. Each detail of the apparatus would require to be separately considered, for a thorough analysis. This is obviously beyond the attempted scope of this paper. The most essential requirements for successful operation may be specified as follows:

Tank Circuits—These should be designed for a kilovoltage value at least 20 times the rated output of the driving tubes in kilowatts. This insures stability of operation in that the tubes readily deliver their energy to the tank circuits in preference to a parasitic circuit. Parasitic circuits are ever present at these "super frequencies."

Stabilizing Circuits. These should be designed with the shortest permissible leads and with the idea of reducing to a minimum the kilovoltage in these circuits.

Symmetry. The circuits must have mechanical and electrical symmetry, particularly in the push-pull type of amplifier to insure that the load is equally shared by the respective tubes.

Grid Leads. Tubes must be provided with extra heavy grid leads and seals due to the heavy charging currents that flow at these frequencies in the inter-electrode capacities. A value of 60 amperes in the grid lead of a water-cooled tube is not uncommon. The inter-electrode capacities of the tubes used must also be reduced to minimum practical values.

Stabilization. The success of any system of cascade amplifications depends upon suppression of regeneration in

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any or all of the stages. There are various means of stabilization one of which in the case of a push-pull amplifier is shown in Figure 7. It will be noted that the potential differences between grid and plate elements of the tubes are reduced by the adjustable balancing condensers connected as shown. The respective potentials of the various parts of the circuit at any instant are as indicated. In practice the

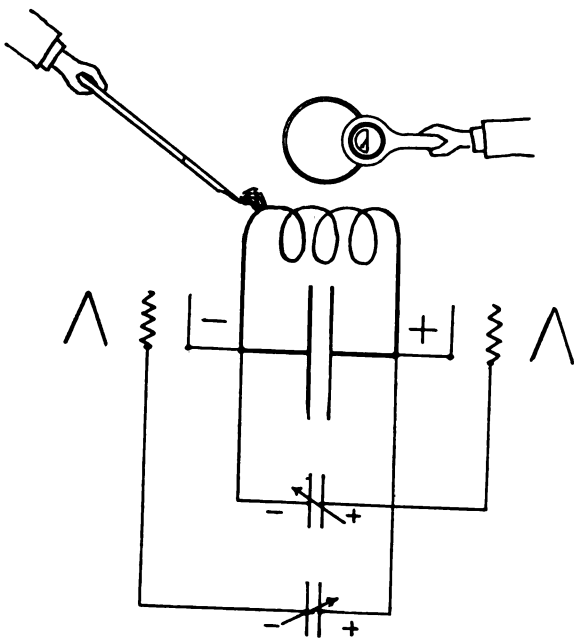

Figure 7—Capacity bridge method of neutralization, showing also application of pick-up loop and stick voltmeter.

process of stabilization adjustment is to excite the grids of the power amplifier with plate voltage off. By means of a pick-up loop meter, the power amplifier tank circuit is adjusted to resonance with the grid excitation. Resonance is indicated by a maximum reading of the loop meter coupled to the tank circuit. The process of stabilization is then to vary the stabilizing condensers until the pick-up loop meter indicates a minimum. These combined adjustments are therefore equivalent to the condition of tank circuit in tune with grid excitation and regeneration at the lowest value obtainable from the existing physical condi-

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tions of the circuits. Figure 7 also indicates the use of the "stick voltmeter". This useful device consists merely of a piece of pointed metal on the end of an insulating rod. The relative potentials of any parts of the circuit are gauged by the lengths of the respective arc discharges. This is a very simple but effective implement for the manipulation of high power short wave circuits.

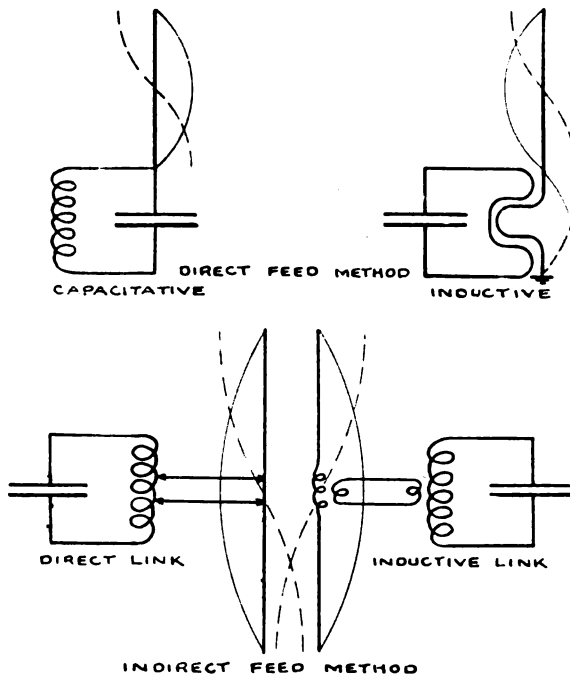

Figure 8—Methods of energizing S. W. Antennas.

V—Antenna and Field Systems

With the advent of short wave transmission, the conventional forms of long wave antennas, which have involved quarter wave radiators in various combinations, have of necessity required radical modification. The underlying reason for this change of point of view is the fact that a relatively short antenna wire can be made several quarter wavelengths long. The result of this situation has been that radiators, in some instances very many wavelengths long, have been used. Short wave antennas are usually referred

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to as ½ wave, ¾ wave, full wave, 7/4 wave antennas, etc. They divide themselves normally into two groups as follows:

(a) Even harmonic antennas.

(b) Odd harmonic antennas.

Since these antennas are normally referred to as odd or even harmonic, and since the harmonic is considered as

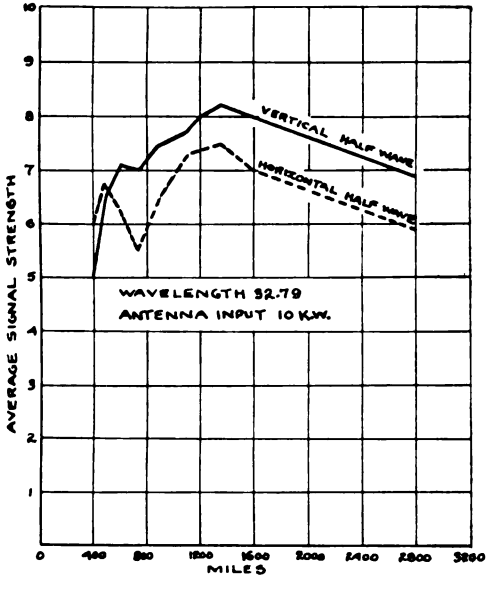

Figure 9—Comparison of signals from half wave vertical and horizontal antennas.

South Schenectady Tests.

a quarter of one wavelength an important relation may be immediately established, namely: the relation between the transmitting wave in meters, and the physical equivalent wire length of the harmonic in feet. This relation is as follows: $\frac{\lambda}{4} = .82\ \lambda$ feet length of wire.

In other words, the length of antenna wire in feet required for one quarter wavelength is equal to 82% of the wavelength in meters. This is on the assumption that the speed of wave propagation in the wire is the same as the speed of light which assumption is accurate enough for all practical purposes. From the above relation the actual physical length of any odd or even harmonic antenna may be calculated.

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Figure 8 has been drawn to indicate various methods of energizing the short wave antenna system. The distribution of current and voltage in the antennas at any instant has been indicated by the solid and dotted waves respectively. The adoption of one antenna system or the other is usually determined by local conditions.

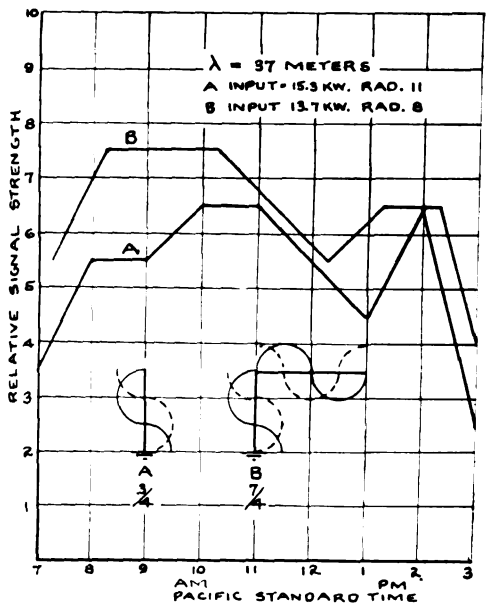

Figure 10—Comparative signal strengths from three-quarter and seven-quarter antennas.
Bolinas, Cal. transmitting, Kokohead, T. H., receiving.

Many tests are being conducted to determine the relative effectiveness of the various antenna types. The radiation effectiveness of any type of short wave radiator is comparatively so high that conclusive results are difficult to obtain. This subject is still being exhaustively investigated. In Figure 9 is shown a comparison obtained at South Schenectady of observed signals from a horizontal and vertical half wave antenna at 32.79 meters and 10 kilowatts antenna input. This data indicates the vertical half wave antenna to be superior to all distances above 500 miles.

Figure 10 plots the comparative strength of signals received at the Radio Corporation of America's station at Kokohead, Hawaii, from a $^3/_4$ and 7/4 wave antenna at the

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Radio Corporation station at Bolinas, California, 2,100 miles distant. This comparison was carefully made by switching from one carefully tuned antenna to the other, alternately at fifteen minute intervals. The approximate forms of the antennas are as indicated at "A" and "B". It will be noted that the 7/4 wave antenna produced a much stronger signal throughout the test. Possible reasons for the greater effectiveness of the longer antenna are:

(a) Greater effective height
(b) Ground losses lessened
(c) Unit antenna stresses lowered.

Conclusive data on antenna systems are difficult to obtain. This subect is continually under investigation by the Radio Corporation and its associated companies. Valuable contributions to this interesting phase of short wave transmission including the effectiveness of directive systems may be expected in the near future.

VI—Static on Short Wave Circuits

The striking advantage of the short wave for long distance communication is the excellent signal-to-static ratio, which is obtained. In fact static ceases to be a problem in the short wave art. Many instances have been noted in which the short wave signal has been entirely readable at the same time that long wave signals were being hopelessly mutilated. Static does exist on certain portions of the short wave band, principally in the spectrum above 30 meters; but it is seldom strong enough to be a practical obstacle to reception. Below 30 meters it may be safely considered as non-existent.

Figure 11 is a graphic record of the relative strengths of signal and static taken at the height of the static season, in July, 1926, on the Radio Corporation of America's New York-Berlin short wave circuits. The relative strength of static on 74 meters as compared to that on 43 meters is of interest. These measurements were made at the receiving station of the Telefunken Company at Geltow, Germany.

VII—Destiny of the Short Wave

It is interesting to speculate on the destiny of short wave communication in the light of present knowledge.

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This speculation of course must be considered solely as the personal judgment of the writer. It assumes that the majority of the defects that now limit the full application of these "super frequencies" have been eliminated. It anticipates by some years the results of present research. Their destiny then may be anticipated as follows:

(a) To spread out the available signaling spectrum to enormous values.

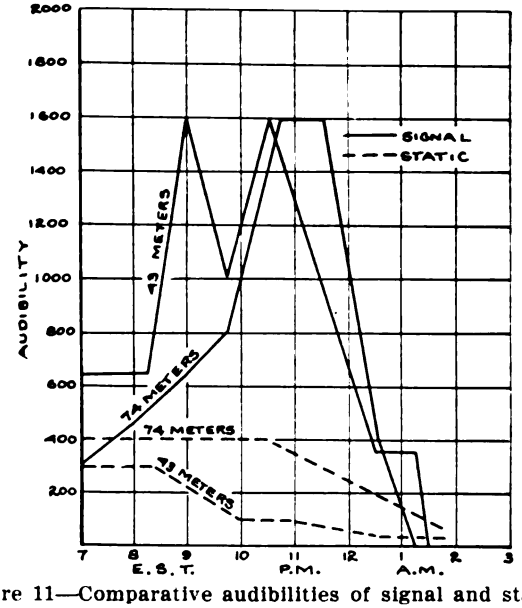

Figure 11—Comparative audibilities of signal and static. New York-Berlin circuit, July, 1926.

(b) To facilitate communication to great distances.

(c) To eradicate the static problem, especially in the tropics.

(d) To permit the fullest development of beam and directive radiation systems.

(e) To permit keying speeds far in excess of present accepted limits.

(f) To make possible the transmission of perfect facsimiles, heralding the end of telegraphy as we know it today.

(g) To form the inevitable background upon which intercontinental television will eventually be built.

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VIII—ACKNOWLEDGMENT

The data which have been presented above obviously represent the work of a group of engineers of the radio Corporation of America, which has been engaged in the design of apparatus and the development of short wave communication under the direct supervision of Mr. C. H. Taylor, Chief Communications Engineer. The helpful suggestions of Messrs. E. V. Amy, H. H. Beverage, C. W. Hansell, F. H. Kroger, and C. W. Latimer, are particularly acknowledged and appreciated.

Summary: Propagation data over the frequency range of 3,000 to 30,000 kilocycles is submitted. A co-relation is shown between wave frequency and angle of projection of the wave front. The effect of ionization on the angle of projection is indicated. Some calculations are given of probable values of attenuation constant.

The importance of frequency stabilization is discussed. Three typical circuits for utilizing control crystals are described.

Features of the design and adjustment of a 20-kw. power amplifier are outlined.

Antenna and antenna feed systems are discussed. Graphical results of comparisons of various antenna types are given.

The relative importance of static at short wave lengths is considered.

The author's anticipation of the destiny of the short wave is summarized.

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MAXIMIZATION METHODS
for
FUNCTIONS OF A COMPLEX VARIABLE

By
Walter Van B. Roberts

(Radio Corporation of America, Technical and Test Department, New York)

In a previous paper it was shown that maximization of the absolute value of a function of a complex variable could be obtained, in some cases very easily, by proper use of the derivative of the function. It will now be shown that there are a number of different conditions that may be imposed upon the derivative, with various results.

Figure 1 shows, in the complex plane, a curve representing the locus of a function of a complex variable as the independent variable traces a certain path (not shown).

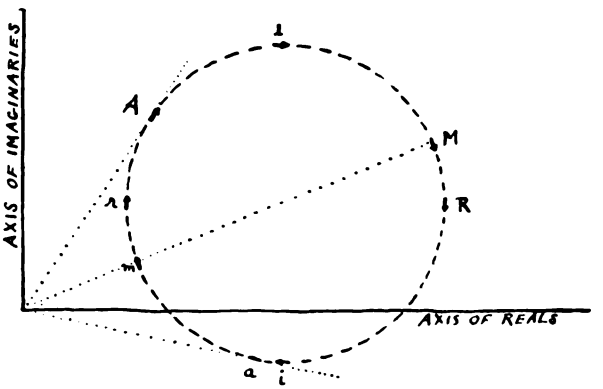

Figure 1

When the independent variable moves an infinitesimal amount dz the dependent variable F moves an amount dF which is an infinitesimal vector forming part of the curve of Figure 1. Furthermore, $dF = \frac{dF}{dz}\, dz$.

Five conditions immediately suggest themselves to be imposed upon dF:

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(1) Require dF to be parallel to the axis of real quantities by equating to zero the imaginary part of $\frac{dF}{dz}dz$ (in other words, argument of $\frac{dF}{dz}dz = 0°$ or $180°$. The solution of this equation gives the value of z that makes the imaginary part of F stationary in value, usually a maximum or a minimum as at "I" or "i" of Figure 1.

(2) Require dF to be parallel to the axis of imaginaries by equating to zero the real part of $\frac{dF}{dz}dz$ (in other words, $arg\frac{dF}{dz}dz = \pm 90°$). The solution of this equation gives the value of z that makes the real part of F stationary in value, usually a maximum or a minimum as at "R" or "r" of Figure 1.

(3) Require dF to be parallel to F by equating to zero the imaginary part of $\frac{\frac{dF}{dz}dz}{F}$. The solution of this equation gives the value of z that makes the argument of F stationary in value, usually a maximum or a minimum as at "A" or "a" of Figure 1.

(4) Require dF to be perpendicular to F by equating to zero the real part of $\frac{\frac{dF}{dz}dz}{F}$. The solution of this equation gives the value of z that makes the absolute value of F stationary in value, usually a maximum or a minimum as at "M" or "m" of Figure 1.

(5) Require dF to vanish by equating to zero both the real and imaginary parts of $\frac{dF}{dz}$. The solutions of these two equations give the conditions under which F is momentarily entirely unaffected by a small change in z. This state of affairs is not shown in Fig. 1.

Following are examples to illustrate the application of the above five processes, the examples being so chosen that the correctness of the results is easily checked by other methods. The examples are worked out at somewhat greater length than would be necessary in actual practice.

(1) Suppose it is desired to determine what adjustment

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of the condenser in Figure 2 will make the reactance of the divided circuit a maximum or minimum. Calling the impedances of the two branches Z_L and Z_C, the value of F is $\frac{Z_L Z_C}{Z_L + Z_C}$ and the independent variable is Z_C. Following rule (1) equate to zero the imaginary part of $\frac{Z_L^2}{(Z_L + Z_C)^2}\, dz_c$. But since the *change* in Z_C is purely imaginary (the condenser being free from losses) the equation may be written: zero equals the real part of $\left(\frac{Z_L}{Z_L + Z_C}\right)^2$. For the purpose of this particular problem it is more convenient to express this equation in the alternative manner: $arg \left(\frac{Z_L}{Z_L + Z_C}\right)^2 = \pm 90°$. From this it follows that $arg \left(\frac{Z_L + Z_C}{Z_L}\right)^2 = \mp 90°$, whence $arg \left(\frac{Z_L + Z_C}{Z_L}\right) = \mp 45°$ or $\pm 135°$ which may be written $arg\ (Z_L + Z_C) = arg\ Z_L \mp 45°$, *or* $= arg\ Z_L \pm 135°$. But of these four possibilities only two make the real part of $(Z_L + Z_C)$ positive, as it must be if the circuit resistances are positive. Therefore the only two conditions to be considered are $arg\ (Z_L + Z_C) = arg\ Z_L - 135°$ and $arg(Z_L + Z_C) = arg\ Z_L - 45°$. If Z_C is so adjusted as to satisfy the first of these the reactance of the divided circuit will be at its maximum value, while if Z_C is so adjusted as to satisfy the second condition the reactance will be a minimum—that is, in this circuit the reactance will have its greatest negative value. It is of interest to note from the above two conditions that if $arg\ Z_L$ is less than $45°$ no positive maximum can occur. Usually however such circuits as that of Fig. 2 have rather low resistances, and in this case greatest positive reactance and greatest negative reactance occur when Z_C is so adjusted as to make the reactance measured around the closed circuit approximately equal to the resistance. If this reactance measured around the circuit is condensive, the reactance of the divided circuit is inductive, and vice versa.

(2) Let a voltage e be applied to an impedance z. The resulting current is $\frac{e}{z}$ and the power absorbed is the real part of the product of voltage and current. In this case the

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complex function F is $\frac{e^2}{z}$ and to find out under what condition its real part is maximum apply rule (2), which leads to the equation: zero equals the real part of $\left(\frac{-e^2}{Z^2}dz\right)$. Before going any further we must decide in what manner z is to be varied. Two simple cases will be considered, (a) the imaginary part of z is kept constant while the real part is varied, and (b) the real part of z is kept constant while the imaginary part is varied. In the first case dz is a real quantity, so the condition becomes: zero equals the real part of $-\frac{e^2}{Z^2}$. This is obviously satisfied if *arg* z is $\pm$ *45°* or $\pm$ *135°*. In other words, for maximum power the resistance should

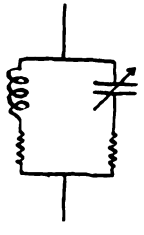

Figure 2

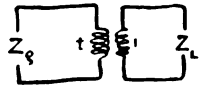

Figure 3

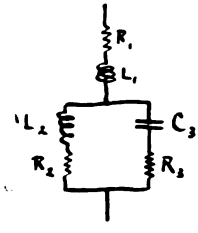

Figure 4

be adjusted to equal the reactance. In case (b) dz is imaginary, hence $-\frac{e^2}{Z^2}$ is real. This condition is satisfied if *arg* z equals *0, 180,* or $\pm$*90*. In other words the power is maximum if the reactance component of z is made zero, and minimum only if the reactance becomes infinite.

(3) Suppose that in the circuit of Fig. 2 it is desired to determine the condenser setting that makes the angle of lead or lag of the divided circuit the greatest. That is, when will *arg* F be maximum or minimum where F is $\frac{Z_L Z_C}{Z_L + Z_C}$? Applying rule 3 we get the equation: zero equals the imaginary part of $\frac{Z_L^2}{(Z_L + Z_C)^2}dz_C \div \frac{Z_L Z_C}{Z_L + Z_C}$. But since dz_C is purely imaginary this is readily reduced to *arg* $\frac{Z_L}{Z_C}\left(\frac{1}{Z_L + Z_C}\right)$ $= \pm$ *90°* whence *arg* $(Z_L + Z_C) =$ *arg* $Z_L -$ *arg* $Z_C \pm$ *90°*. If the circuit resistances are positive only the latter alterna-

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tive may be used. Hence the condenser must be so adjusted that if a voltage were inserted in the closed circuit a current would be produced that would lag the voltage by an angle as many degrees less than *90* as the sum of the angles by which Z_L and Z_C fail to be pure reactances. For example if $arg\ Z_L = 85°$ and $arg\ Z_C$ is between *89°* and *90°* then $arg\ (Z_L + Z_C)$ must lie between *84°* and *85°*. That is, the condenser should be so adjusted that current in the closed circuit would lag voltage by between 84 and 85 degrees. This gives the value of capacity fairly closely so that $arg\ Z_C$ may be estimated more closely than originally assumed, and using this more accurate value the adjustment of the condenser may be obtained still more closely, and so on till as good an approximation as desired is attained. The interesting thing about this problem is that for values of capacity between zero and infinity there is only a minimum of the argument, and this corresponds usually to a condensive reactance of the divided circuit.

(4) Fig. 3 shows a generator of internal impedance Z_g connected through an ideal transformer (that is, one having no losses or leakage, and infinite reactance windings) of turns ratio to a load of impedance Z_L. The problem is to find what turns ratio will result in the greatest load current. The expression for load current is easily found to be $\frac{t\,e}{Z_g + t^2 Z_L}$, and this is the function F to be maximized. Following rule 4, set zero equal to the real part of $\frac{Z_g - t^2 Z_L\ d\,t}{Z_g + t^2 Z_L\ \ t}$ It is readily shown, either geometrically or by substitution of assumed real and imaginary parts for Z_g and Z_L, that this equation is satisfied by $t^2 = \left|\frac{Z_g}{Z_L}\right|$ which is a well known rule.

(5) The impedance of the circuit element shown in Fig. 4 is $Z_1 + \frac{Z_2 Z_3}{Z_2 + Z_3}$. Call this impedance F and apply rule 5, considering frequency as the independent variable. The resulting equation is $0 = \frac{dz_1}{df} + \frac{Z_2 \frac{dz_2}{df} + Z_3 \frac{dz_2}{df}}{(Z_2 + Z_3)^2}$. It is easy to show that if $R_2 = R_3$ and $f^2 = \frac{1}{4\pi^2 L_2 C_3}$ the real part of the second

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term is zero, and the imaginary part becomes $j2\,L_2\left(R^2-\frac{L_2}{C_3}\right)\frac{1}{4R^2}$. So in order to satisfy the equation it is only necessary to add the condition $L_1=2L_2\left[\frac{L_2}{C_3}-R^2\right]\frac{1}{4R^2}$ An interesting case occurs when $L_1=0$ and $R^2=\frac{L_2}{C_3}$ for in this case the derivative is zero for all values of frequency, so that the impedance of the combination is completely independent of frequency. This case is only one of several of the sort.

In conclusion, it should be noted that no test has been given for determining whether or not it is legitimate to differentiate a given function of a complex by the same formal process as used for functions of real variables. The reason for omitting so fundamental a necessity is that the test may be found in any text book on complex functions, and also because any function obtained from circuit theory will almost surely be analytic, which means that it can be differentiated exactly as though all the letters represented real quantities.

SUMMARY—The maxima and minima of a function of a real variable are found by equating to zero the derivative of the function. In the case of a function of a complex variable however the derivative is a vector quantity, so that conditions may be imposed upon its direction as well as upon its magnitude. These various conditions lead to maxima and minima of the various aspects of the function. Rules are developed for setting up equations giving the various maximizing conditions, and a simple example is given illustrative of the use of each rule.

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A MATHEMATICAL STUDY OF RADIO FREQUENCY AMPLIFICATION*

BY
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INTRODUCTION

There are two general methods of radio frequency amplification, tuned and untuned. The latter is seldom used except in super-heterodyne circuits in conjunction with one or more tuned units or a band pass filter to provide the selectivity. Theoretically, this untuned super-heterodyne with a band pass filter is the best type possible. However, the voltage amplification of the untuned type of transformer is at present very unsatisfactory, as tube and winding capacities are exceedingly troublesome.

There are two methods of tuning in use. We may either tune the frequency to the circuits or the circuits to the frequency. The first is the super-heterodyne method and the second is the common method of the neutrodyne and similar circuits.

It is this latter type that we shall study particularly though many of our results are more general than this.

Throughout we shall assume a steady state, leaving transients for further consideration. By this assumption we may apply the ordinary vector methods of alternating current theory.

THE AIR CORE TRANSFORMER

The fundamental circuit of the transformer is that shown in Figure 1, where the symbols have their usual meaning.

In terms of vectors and complex operators† we may

*Presented before the Canadian Section at Toronto, January 5, 1927.

†All vectors and complex operators are printed in Clarendon type, scalars are in ordinary type. If a distinguishing mark is required in script, a bar or dot over the letter will provide it. This is usually unnecessary.

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equate the emfs. and impedance drops around each circuit. We get

$$\mathbf{E}_1 = \mathbf{Z}_1 \mathbf{I}_1 - \mathbf{m}\,\mathbf{I}_2 \tag{1}$$

$$0 = \mathbf{Z}_2 \mathbf{I}_2 - \mathbf{m}\,\mathbf{I}_1 \tag{2}$$

where $\mathbf{Z}_1 = R_1 + X_1 j, X_1 = \left(L_1\omega - \frac{1}{C_1\omega}\right)$, $\mathbf{m} = M\omega j$, etc.

Eliminating $\mathbf{I}_1$ and solving for $\mathbf{I}_2$ we get

$$\mathbf{I}_2 = \frac{\mathbf{E}_1}{\frac{\mathbf{Z}_1\,\mathbf{Z}_2}{\mathbf{m}} - \mathbf{m}}\,. \tag{3}$$

This is a general vector equation for the secondary current which we shall have to study to find out how to make it a

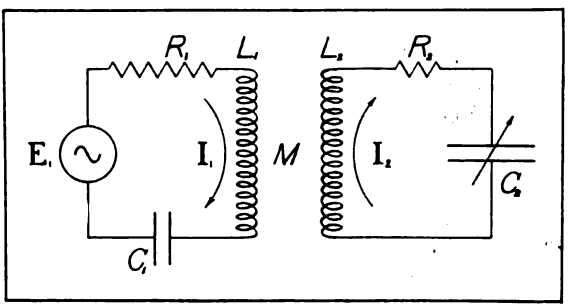

Figure 1—The Fundamental Circuit of the Transformer

maximum. We could continue to work with vectors' but it is better to change to absolute values. We have

$$\mathbf{Z}_1\,\mathbf{Z}_2 = (R_1 + X_1 j)(R_2 + X_2 j) = a + bj, \tag{4}$$

where $a = R_1 R_2 - X_1 X_2$,

and $b = R_1 X_2 + R_2 X_1$.

Then (3) becomes

$$\mathbf{I}_2 = \frac{\mathbf{E}_1}{\frac{a+bj}{M\omega j} - M\omega j} = \frac{\mathbf{E}_1}{\frac{b}{M\omega} - \left(\frac{a}{M\omega} + M\omega\right)j}\,. \tag{5}$$

If $K\mathbf{I}_2$ is the conjugate of $\mathbf{I}_2$,

$$I_2^2 = \mathbf{I}_2\,K\mathbf{I}_2 = \frac{E_1^2}{\left(\frac{b}{M\omega}\right)^2 + \left(\frac{a}{M\omega} + M\omega\right)^2} = \frac{E_1^2}{M^2\omega^2 + 2a + \frac{a^2+b^2}{M^2\omega^2}}, \tag{6}$$

[1] W. B. Roberts, PROCEEDINGS I. R. E., Vol. 14, No. 5, Oct. 1926, "A Method for Maximization in Circuit Calculation."

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which is the general equation in scalars in place of the vector equation (3).

Conditions for Maximum

(a) If M only varies I_2 is a maximum when

$$M^2\omega^2 = \sqrt{a^2+b^2} = Z_1 Z_2 , \tag{7}$$

as may be seen by inspection of (6) or by differentiating.

(b) If X_2 only varies, I_2 is a maximum when

$$\frac{da}{dX_2} + \frac{1}{M^2\omega^2}\left(a\frac{da}{dX_2} + b\frac{db}{dX_2}\right) = 0 , \tag{8}$$

which, on substituting for (a) and (b), leads to

$$X_2 = \frac{M^2\omega^2}{Z_1^2} X_1 . \tag{9}$$

(c) If X_1 only varies, I_2 is a maximum when

$$X_1 = \frac{M^2\omega^2}{Z_2^2} X_2 . \tag{10}$$

Now these three maximum conditions are not independent. If we satisfy any two of them the third is also satisfied. All three are satisfied if

$$\frac{X_1}{R_1} = \frac{X_2}{R_2} \tag{11}$$

and $$M^2\omega^2 = Z_1 Z_2 = R_1 R_2 + X_1 X_2 = \frac{R_2}{R_1} Z_1^2 . \tag{12}$$

But these two equations can always be satisfied by adjusting any two of X_1, X_2 and M; it is not necessary to vary all three.

The most obvious result of this is that it is unnecessary to tune an aerial which is coupled to a tuned secondary circuit.

Maximum Secondary Current

When equations (11) and (12) hold the general equation (6) reduces to the simple form

$$I_2 = \frac{E_1}{2\sqrt{R_1 R_2}} \tag{13}$$

This is the maximum possible secondary current.

A Tube Constant

A tube acts as a generator with an emf. μE_g where E_g is the grid voltage, and an internal resistance R_p, the plate

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impedance. Then for an interstage transformer we have

$$E_1 = \mu E_g$$

$$R_1 = R_p, \text{ very closely,}$$

and the voltage applied to the next grid is

$$\frac{1}{C_{2\omega}} I_2 = L_{2\omega} I_2, \text{ nearly,}$$

for $L_{2\omega} = \frac{1}{C_{2\omega}}$ nearly in an interstage transformer.

Hence the amplification is, under ideal conditions,

$$\frac{E_{g2}}{E_{g1}} = \frac{1}{2} \cdot \frac{\mu}{\sqrt{R_p}} \cdot \frac{L_{2\omega}}{\sqrt{R_2}} \cdot \tag{14}$$

Now $\frac{\mu}{\sqrt{R_p}}$ is a tube constant* which should be used in place of the mutual conductance as a measure of its efficiency in a radio frequency circuit, while $\frac{L_{2\omega}}{\sqrt{R_2}}$ is a factor of the secondary coil which might well be listed for various coils.

The Primary Current

From (2) and (3) we get

$$\mathbf{I}_1 = \frac{\mathbf{E}_1}{\mathbf{Z}_1 + \frac{M^2 \omega^2}{\mathbf{Z}_2}} \cdot \tag{15}$$

Under ideal conditions, when (11) and (12) hold, this reduces to

$$\mathbf{I}_1 = \frac{\mathbf{E}_1}{2R_1} \cdot \tag{16}$$

Hence we see that the reaction from the secondary current is to make the primary circuit effectively a pure resistance of twice its true value; all primary reactance has in effect disappeared.

The Secondary Current

From equations (3), (11) and (12) we get:

$$\mathbf{I}_2 = \frac{\mathbf{E}_1}{2\sqrt{R_1 R_2}} \cdot \frac{X_1 + R_1 j}{Z_1} \cdot \tag{17}$$

This shows that when the reactance of the primary circuit is small in comparison with its resistance the secondary

*The square of this constant, $\left(\frac{\mu}{R_p}\right)$, is sometimes used; it is then a conductance.

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current leads the primary voltage by nearly a right angle. The departure from a right angle is just enough to provide the reactive voltages required in the primary.

Energy Relations

The power dissipated in each circuit is:

In the primary, $I_1^2 R_1 = \frac{E_1^2}{4 R_1}$, (18)

In the secondary, $I_2^2 R_2 = \frac{E_1^2}{4 R_1}$. (19)

Both are the same, which is the required condition for the maximum output from the primary circuit to the secondary circuit.

Fixed Mutual Inductance

In many circuits the primary reactance and the mutual inductance are fixed, only the secondary reactance is varied. Sometimes the mutual inductance is varied in a rough manner with the secondary condenser. The question of the behaviour of these circuits arises.

The Primary Current

Substituting (9) into (15) we get

$$\mathbf{I}_1 = \mathbf{E}_1 \cdot \frac{\left(R_1 R_2 + \frac{M^2 \omega^2}{Z_1^2} X_1^2\right) - X_1 \left(R_2 - R_1 \frac{M^2 \omega^2}{Z_1^2}\right) j}{M^2 \omega^2 R_1 + Z_1^2 R_2} \cdot \tag{20}$$

which shows that if M is greater than its proper value, the nature of the primary reactance is in effect reversed, an inductive reactance is changed to a condensive reactance and vice versa.

If M is less than its proper value, the nature of the primary reactance is unchanged.

The Secondary Current

Substituting (9) into (3) we find the maximum secondary current when X_2 only is varied to be

$$\mathbf{I}_2 = \frac{\mathbf{E}_1}{\left(\frac{M\omega}{Z_1} R_1 + \frac{Z_1}{M\omega} R_2\right)} \cdot \frac{X_1 + R_1 j}{Z_1} \cdot \tag{21}$$

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Comparing this with (17) we notice that the phase with respect to the primary voltage is the same as under the best conditions. On reducing (21) to its scalar equivalent we get

$$I_2 = \frac{E_1}{\left(\frac{M\omega}{Z_1}R_1 + \frac{Z_1}{M\omega}R_2\right)}. \qquad (22)$$

Using this formula the curves for an aerial type and an interstage type of transformer were drawn. They are shown

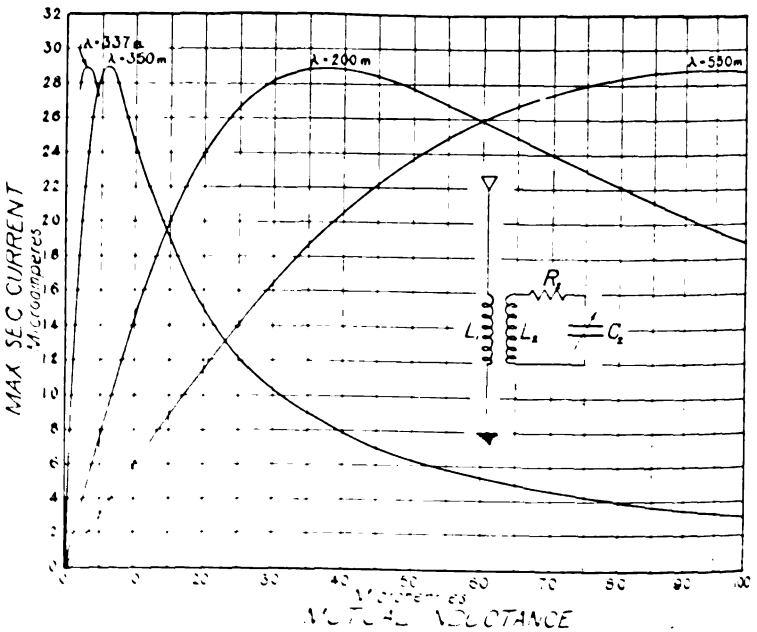

Figure 2. Characteristics of an Aerial Type Transformer when the secondary only is tuned. Three wavelengths, 200, 350 and 550 meters are shown. The constants used are:

R_1 = 30 ohms	R_2 = 10 ohms
L_1 = 2[illegible] microhenrys	L_2 = 290 microhenrys
C_1 = 32[illegible] micromicrofarads	C_2 = 350 micromicrofarads

E_1 = 1 millivolt

Natural Primary Wavelength = 337 meters

in Figures 2 and 3 respectively. For simplicity the resistances are assumed not to vary with the frequency, thus making all three curves have the same peak in each case. An equivalent aerial with lumped constants is assumed.[1]

[1] [illegible] Papers, No. 124, "Electrical [illegible]"

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Figure 2 is extremely interesting as it shows that very loose coupling is required at and around the frequency for which the aerial is naturally resonant. This has been called an absorption band because, with fixed coupling, as is often used, the response is very poor in this region. But it is not

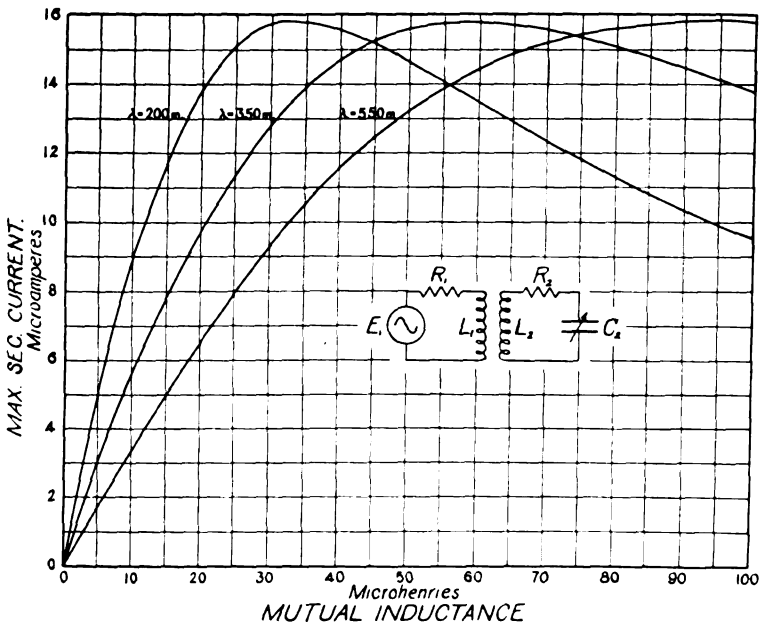

Figure 3—Characteristics of an Interstage Type Transformer when the secondary only is tuned. Three wavelengths, 200, 350 and 550 are shown. The constants used are:

$R_1 = 10{,}000$ ohms $\quad R_2 = 10$ ohms

$L_1 = 46$ microhenrys $\quad L_2 = 290$ microhenrys

$C_1 = \infty \quad C_2 = 350$ micromicrofarads max.

$E_1 = 10$ millivolts

an absorption band in any sense of the word, it is merely a band of frequencies in which the coupling must be very accurately adjusted to small values.

This band may be avoided by using a series condenser in the aerial so as to make its natural frequency higher than any in range over which the set must operate. A loading coil will also have the same effect by decreasing the natural frequency below the lowest of the range.

But in case some one local station is troublesome we may use this loose-coupling band as a means of getting

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greater selectivity in the region of that station by making our aerial naturally resonant to that station. This almost seems contrary to common sense at first.

Figure 3 shows that the coupling of all transformers should be varied with the frequency but that in the interstage transformer the high primary resistance makes the adjustment very rough at all frequencies.

Tube Capacities

In all radio amplification the capacities between tube elements are troublesome. Various methods have been devised for neutralizing the grid to plate capacity to prevent

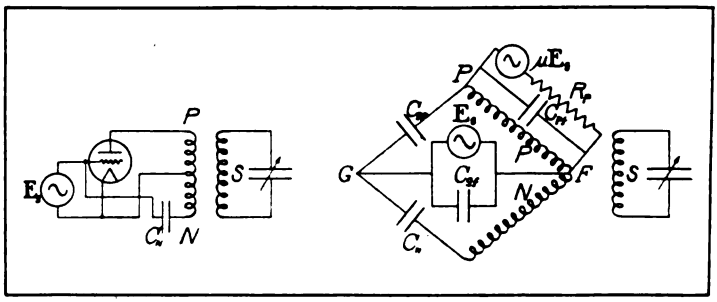

Figure 4—The Roberts' Circuit and its Equivalent Bridge Form.

all interaction between the grid circuit and the plate circuit other than the true electronic action of the tube. All of these methods are simply inductance capacity bridges, or perhaps just capacity bridges, some in the grid circuit, some in the plate.

The Roberts' circuit and its equivalent bridge arrangement are shown in Figure 4. This bridge is not quite perfect, for the primary coil P has some resistance, whereas it should have none at all, and the coefficient of coupling between N and P is never quite unity as it should be.

Note that the primary and the neutralizing coil should be quite closely coupled and *must* be related to the secondary in a symmetrical manner, for the secondary has a greater effect upon the primary emf. than the primary itself.

We also see that the tube has an input capacity of nearly $(C_{gf} + 2C_{gp})$, while there is an output capacity shunting the plate to the filament which is effectively $(C_{pf} + C_{gp})$. It is at least C_{pf} in any circuit.

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In case any should think this negligible just note that at 1,500 kilocycles, 10 micromicrofarads had an impedance of 10,600 ohms, which is of the same order as the plate impedance of many tubes.

Output Capacity

Let us call the output capacity C_o, which must include all capacities from plate to filament. The equivalent circuit is shown in Figure 5.

Let the voltage from "a" to "b" $= \mathbf{V}_x$.

Then we may write the equations

$$\mathbf{V}_x = \mathbf{E}_1 - R_p(\mathbf{I}_o + \mathbf{I}_1), \tag{23}$$

$$\mathbf{V}_x = \mathbf{Z}_1 \mathbf{I}_1 - \mathbf{m}\mathbf{I}_2, \tag{24}$$

$$\mathbf{I}_o = \mathbf{V}_x C_o \omega j, \tag{25}$$

$$0 = \mathbf{Z}_2 \mathbf{I}_2 - \mathbf{m}\mathbf{I}_1, \tag{26}$$

where $\mathbf{Z}_1$ does not include R_p as it has previously.

Eliminating $\mathbf{V}_x$ and $\mathbf{I}_o$ we get

$$\frac{\mathbf{E}_1}{1 + R_p C_o \omega j} = \left(\mathbf{Z}_1 + \frac{R_p}{1 + R_p C_o \omega j}\right)\mathbf{I}_1 - \mathbf{m}\mathbf{I}_2 \tag{27}$$

$$0 = \mathbf{Z}_2 \mathbf{I}_2 - \mathbf{m}\mathbf{I}_1 \tag{28}$$

From these equations it is seen that the tube behaves exactly as though it had a complex amplification factor

$$\mu = \frac{\mu}{1 + R_p C_o \omega j}, \tag{29}$$

and a complex plate impedance*

$$\mathbf{R}_p = \frac{R_p}{1 + R_p C_o \omega j} = \frac{R_p}{1 + R_p^2 C_o^2 \omega^2} - \frac{R_p^2 C_o \omega}{1 + R_p^2 C_o^2 \omega^2} j \tag{30}$$

which is as though the plate impedance were reduced in the ratio of $\frac{1}{1 + R_p^2 C_o^2 \omega^2}$ and a series capacity $\frac{1 + R_p^2 C_o^2 \omega^2}{R_p^2 C_o \omega}$ were in the plate circuit.

---

*This is merely a special case of a general theorem which says that generators in parallel, having emfs. $\mathbf{E}_1$, $\mathbf{E}_2$, ------$\mathbf{E}_n$, any of which may be zero, and impedances $\mathbf{Z}_1$, $\mathbf{Z}_2$, ------$\mathbf{Z}_n$, are exactly equivalent to one generator with an impedance $\mathbf{Z}$, where $\frac{1}{\mathbf{Z}} = \frac{1}{\mathbf{Z}_1} + \frac{1}{\mathbf{Z}_2} + \text{------} \frac{1}{\mathbf{Z}_n}$ and an emf., $\mathbf{E} = \mathbf{Z}\left(\frac{\mathbf{E}_1}{\mathbf{Z}_1} + \frac{\mathbf{E}_2}{\mathbf{Z}_2} + \text{------} \frac{\mathbf{E}_n}{\mathbf{Z}_n}\right)$. This is easily proved as above by assuming an unknown bus bar voltage.

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The effect of the output capacity upon the secondary current is a slight change in phase and value. If we neglect the resistance and reactance of the primary coil in comparison with the fictitious plate impedance and the reactance of the fictitious series condenser there is no change whatever.

We may prove this by substituting (29) and (30) into (17) or by studying the vector diagram to which these equations lead.

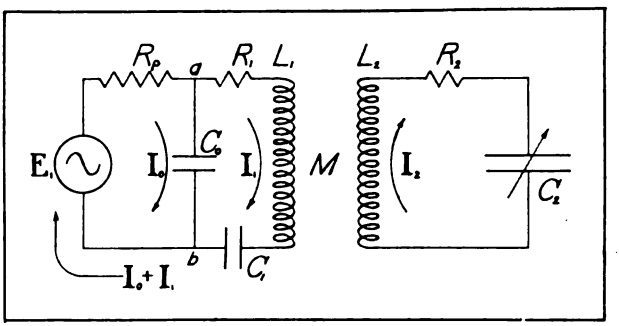

Figure 5—Fundamental Circuit of the Interstage Transformer When the Output Capacity of the Tube is Considered.

However, the output capacity forces the secondary circuit to have a considerable condensive reactance which in turn forces a considerably closer coupling than we should expect if we neglected it, for equations (11) and (12) must be satisfied to get the best results.

To construct this note that

$$\mathbf{E}_1 = \frac{\mu \mathbf{E}_{g1}}{1 + R_p C_o \omega j}, \tag{31}$$

which lags $\mathbf{E}_{g1}$ by an angle $tan^{-1} R_p C_o \omega$

and $$\mathbf{I}_1 = \frac{\mathbf{E}_1}{2\left(R_1 + \frac{R_p}{1 + R_p^2 C_o^2 \omega^2}\right)}, \tag{32}$$

which is in phase with $\mathbf{E}_1$

Then we have

$$\mathbf{E}_2 = M \omega j \mathbf{I}_1, \tag{33}$$

which leads $\mathbf{I}_1$ by $90°$.

Now $\mathbf{I}_2$ leads $\mathbf{E}_2$ by an amount $tan^{-1} \frac{X_2}{R_2} = tan^{-1} \frac{X_1}{R_1}$, which is

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usually only slightly less than $tan^{-1} R_p C_o \omega$. This brings $\mathbf{I}_2$ nearly at right angles to E_{g1} and its value is given by

$$I_2 = \frac{E_1}{2\sqrt{\left(R_1 + \frac{R_p}{1 + R_p^2 C_o^2 \omega^2}\right) \cdot R_2}} \cdot \qquad (34)$$

Finally the condenser voltage E_{g2} lags nearly $90°$ behind $\mathbf{I}_2$ and is given by

$$E_{g2} = \frac{\mathbf{I}_2}{C_2 \omega j} \cdot \qquad (35)$$

which completes the diagram.

Other vectors can be added as desired.

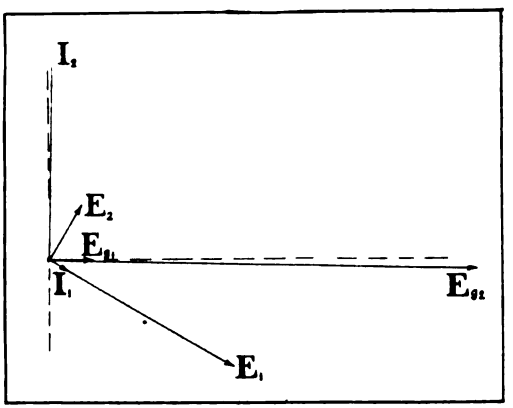

Figure 6—A Vector Diagram of the Interstage Type Transformer When the output capacity is considered.

REGENERATION

Strictly speaking, the regenerative transformer also comes under the heading of radio frequency amplification. However, it is far too lengthy and important a subject to be added here, as it requires to be treated from first principles and not as an extension of this work.

MUTUAL CAPACITY

We have throughout assumed that the capacity between the primary and the secondary is negligible. A very little thought will show that all such capacity is very objectionable, especially as the two coils are usually connected to-

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gether at one end by a common battery. Such capacity must be kept as small as possible.

SUMMARY: The necessary conditions for the maximum secondary current are $\frac{X_1}{R_1} = \frac{X_2}{R_2}$ and $M^2\omega^2 = Z_1 Z_2$ which may be established with only two variables making it unnecessary to tune the primary. The secondary current is then

$$I_2 = \frac{E_1}{2\sqrt{R_1 R_2}}$$

This leads to a tube constant $\frac{\mu}{\sqrt{R_p}}$ as a measure of its efficiency in a R. F. amplifier, also a coil constant $\frac{L\omega}{\sqrt{R}}$ for the secondary coil efficiency.

When the secondary only is tuned the maximum secondary current occurs when $X_2 = \frac{M^2\omega^2}{Z_1^2} X_1$. Curves of this maximum secondary current are given in Figures 2 and 3 and discussed there.

The effect of tube output capacity is then studied, the solution being made to depend upon the previous work. In fact all additional problems such as combined impedance and mutual inductance coupling, mutual capacity between the windings, etc., can be made to depend upon the fundamental solution in a similar manner. We find that the effect of the output capacity is the same as though the tube had a complex amplification factor and a complex plate impedance.

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DISCUSSION ON VACUUM TUBE NOMENCLATURE* (CHAFFEE)

M. W. Arps: While the writer agrees with Mr. Metcalfe that a nomenclature suitable for printing at moderate cost is desirable, it is desired to take exception to his estimate of the number of readers of the Institute interested in a paper of this kind.

Without a nomenclature of the extent of Dr. Chaffee's it would not be possible to write out the equations for the solution of certain vacuum tube problems, a fair example of of which was Dr. Chaffee's article on "Detection" in Volume 15 Number 2 (February, 1925) Proceedings.

The value of articles of this nature are not in the number of readers that are immediately interested, but in the number of members of the Institute who will have occasion to use this article in the next five years, and will have the paper available when they do become interested in this subject.

It is not believed anyone would deliberately set down and calculate the efficiency of a detector by the formulas as given. On the other hand, after my engineer had found by test that one vacuum tube was better than the average, it is believed there would be no more exact, or quicker method of locating the reason the tube was better, than to measure the constants of the tube, drop them into Dr. Chaffee's formulas, and grind out the result.

Even though they never read the paper through until the need arises, the fact that an engineer engaged in radio consulting work may be called on at almost any time to handle a problem of this nature is what makes papers of this kind of value to members of the Institute.

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*Proc. I. R. E., Vol. 15, No. 3, March 1927, p. 182, discussion p. 253.

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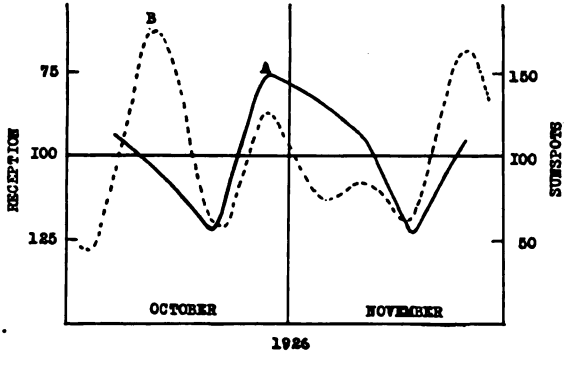
DISCUSSION ON LONG DISTANCE RECEIVING MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1925 (L. W. AUSTIN)

Greenleaf W. Pickard: Mr. Sreenivasan's discussion[1] of Dr. Austin's Paper contains a short tabulation of weekly averages of Bangalore reception of Madras Radio, the distance being 295 kilometers and the frequency 75 kilocycles. Although the period of this tabulation is too short for a good comparison with terrestrial and cosmic elements, it may be of interest to contrast these measurements with sunspots for the same interval. Taking his weekly averages



as ratios with the mean of the entire series, and plotting with inverted ordinates, the full-line curve A is obtained. The dotted-line curve B is made from the daily ratios of moving 7 and 27 day averages of the Wolfer Provisional Sunspot Numbers.

These curves suggest an inverse relation between Indian reception measurements and solar activity, such as that which I have found[2] for Dr. Austin's day reception at Washington from distant low frequency stations during 1926. It

[1] Proceedings of the Institute of Radio Engineers, Vol. 15, No. 2, page 155, February, 1927.

[2] Proceedings of the Institute of Radio Engineers, Vol. 15, No. 2, Fig. 4, page 93, February, 1927

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would be very interesting if Mr. Sreenivasan would compare, by the method which I have indicated above, a longer series of his 1926 measurements with sunspots, to determine whether this apparent inverse relation persists throughout the year.

In view of an investigation now in progress, in which I have already analyzed Dr. Austin's reception data for 1917, 1922 and 1923, finding in each of these years a direct instead of an inverse relation between sunspots and day reception, it would seem that 1926 might in this regard be an exception to a general rule.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued Feb. 15, 1927—March 29, 1927

By

JOHN B. BRADY

Patent Lawyer, Ouray Building, Washington, D. C.

1,617,618—ELECTRIC CONDENSER—H. A. DOUGLAS, of Bronson, Mich. Filed June 16, 1924, issued Feb. 15, 1927.

1,617,467—RADIO SIGNALING SYSTEM—L. S. UPHOFF, of Brooklyn, N. Y. Filed Dec. 30, 1922, issued Feb. 15, 1927. Assigned to American Telephone & Telegraph Co.

1,617,956—CRYSTAL DETECTOR—L. LANE, of Habana, Cuba. Filed March 10, 1925, issued Feb. 15, 1927.

1,617,974—RADIO SIGNALING SYSTEM—W. C. WHITE, of Schenectady, N. Y. Filed Aug. 23, 1915, issued Feb. 15, 1927. Assigned to General Electric Co.

1,618,017—F. LOWENSTEIN, of New York. Filed June 28, 1910, issued Feb. 15, 1927. Assigned to Radio Corporation of America.

1,618,018—F. LOWENSTEIN, of Rockaway Park, N. Y. Filed Feb. 25, 1915, issued Feb. 15, 1927. Assigned to Radio Corporation of America.

1,618,255—VARIABLE NEUTRALIZING CONDENSER—JOHN J. WOOD, Denver, Colo. Filed May 16, 1925, issued Feb. 22, 1927.

1,618,298—VALVE AMPLIFIER—HANS E. RUKOP, Berlin, Germany. Filed Sept. 3, 1921, issued Feb. 22, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie.

1,618,299—AERIAL SYSTEM EMPLOYED IN RADIO SIGNALING—S. B. SMITH and G. M. WRIGHT, Chesterfield, England. Filed Mar. 8, 1921, issued Feb. 22, 1927. Assigned to Radio Corp. of America.

1,618,399—VARIABLE CONDENSER—R. M. WISE and E. G. DANIELSON, San Francisco, Cal. Filed July 26, 1924, issued Feb. 22, 1927. Assigned to E. T. Cunningham.

1,618,499—ELECTRICAL APPARATUS—C. P. WHITE, Lynn, Mass. Filed Nov. 6, 1923, issued Feb. 22, 1927.

1,618,468—RADIO RECEIVING EQUIPMENT—J. C. R. PALMER, New Rochelle, N. Y. Filed June 13, 1923, issued Feb. 22, 1927. Assigned to Western Electric Co.

1,618,816—LIGHT CONTROL SWITCH—H. R. DAVIES, Valley Cottage, N. Y. Filed May 3, 1918, issued Feb. 22, 1927. Assigned to Industrial Research Corp.

1,618,974—VARIABLE CONDENSER—C. N. CAHUSAC, Newark, N. J. Filed Apr. 16, 1925, issued Mar. 1, 1927. Assigned to U. S. Tool Co. Inc.

1,619,201—CONDENSER—J. A. FRIED, Flushing, N. Y. Filed Apr. 29, 1926, issued Mar. 1, 1927. Assigned to Dubilier Condenser Corp.

1,619,223—CONDENSER—H. R. VAN DEVENTER, New York City. Filed May 17, 1924, issued Mar. 1, 1927. Assigned to Dubilier Condenser Corp.

1,619,318—ELECTROSTATIC DETECTOR AND AMPLIFIER—F. E. SUMMERS, Near Memphis, Mo. Filed Jan. 19, 1920, issued Mar. 1, 1927. Assigned to Magnavox Co.

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1,619,544—RADIO APPARATUS—A. F. SNAFORD, Knoxville, Tenn. Filed Aug. 21, 1925, issued Mar. 1, 1927.

1,619,318—ELECTROSTATIC DETECTOR AND AMPLIFIER—F. E. SUMMERS, near Memphis, Mo. Filed Jan. 19, 1920, issued Mar. 1, 1927. Assigned to Magnavox Co.

1,619,545—RADIO APPARATUS—A. F. SANFORD, Knoxville, Tenn. Filed Sept. 23, 1925, issued Mar. 1, 1927.

1,619,597—ELECTRICAL CONDENSER—M. CLARK, Chicago, Ill. Filed Feb. 1, 1926, issued Mar. 1, 1927.

1,619,892—MEANS FOR TRANSFERRING HIGH FREQUENCY ENERGY—R. S. OHL, New York City. Filed Dec. 4, 1924, issued Mar. 8, 1927. Assigned to American Telephone & Telegraph Co.

1,619,974—VALVE OPERATING MECHANISM—G. E. A. HALLETT, Dayton, O. Filed Nov. 24, 1924, issued Mar. 8, 1927. Assigned to General Motors Research Corp.

1,620,020—VARIABLE CONDENSER—C. HARDY, New York City. Filed Dec. 10, 1923, issued Mar. 8, 1927. Assigned to Amsco Products, Inc.

1,620,204—FREQUENCY INDICATION AND CONTROL—R. A. HEISING, Millburn, N. J. Filed Dec. 29, 1924, issued Mar. 8, 1927. Assigned to Western Electric Co. Inc.

1,620,231—RADIO ANTENNA SYSTEM—D. W. RICHARDSON, Princeton, N. J. Filed Apr. 26, 1922, issued Mar. 1, 1927.

1,620,244—VARIABLE CONDENSER—F. E. SWOPE, JR., Philadelphia, Pa. Filed May 18, 1923, issued Mar. 8, 1927.

1,620,318—HIGH FREQUENCY ARC CONVERTER—R. R. BEAL and H. F. ELLIOTT, Palo Alto, Cal. Filed Jan. 9, 1922, issued Mar. 8, 1927. Assigned to Federal Telegraph Co.

1,620,655—DIRECTIVE RADIO RECEIVING SYSTEM—R. A. HEISING, East Orange, N. J. Filed Sept. 8, 1920, issued Mar. 15, 1927. Assigned to Western Electric Co.

1,620,656—CARRIER WAVE SIGNALING SYSTEM—R. A. HEISING, Millburn, N. J. Filed Dec. 1, 1924, issued Mar. 15, 1927. Assigned to Western Electric Co., Inc.

1,620,973—RADIO RECEIGING CIRCUIT—W. KAUFMAN, Paterson, N. J. Filed Apr. 14, 1923, issued Mar. 15, 1927.

1,621,034—RADIO RECEIVING SYSTEM—J. SLEPIAN, Swissvale, Pa. Filed Jan. 13, 1922, issued Mar. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,621,037—RECEIVING SYSTEM FOR HIGH FREQUENCY ELECTRICAL OSCILLATIONS—A. H. TAYLOR and L. C. YOUNG, Washington, D. C. Filed Jan. 19, 1924, issued Mar. 15, 1927.

1,621,058—ELECTRIC CONDENSER—O. BURGER, Berlin, Charlottenburg, Germany. Filed Feb. 10, 1925, issued Mar. 15, 1926. Assigned to Westinghouse Electric & Mfg. Co.

1,621,062—RADIO SYSTEM—L. W. CHUBB, Edgewood Park, Pa. Filed June 17, 1920, issued Mar. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,621,253—VARIABLE ELECTRIC CONDENSER, VARIOMETER AND THE LIKE—A. F. HOLLIDAY, Wigan, England. Filed Dec. 8, 1924, issued Mar. 15, 1927.

1,621,379—VARIABLE CONDENSER—T. T. RODGERS, M. A. GIBLIN, F. K. MOORE and G. T. MOORE, Milwaukee, Wis. Filed Sept. 24, 1924, issued Mar. 15, 1927. Assigned to Rodgers Radio Corp.

1,621,659—CONDENSER—J. B. DEMPSTER, Iowa City, Ia. Filed Dec. 14, 1922, issued Mar. 22, 1927.

1,621,721—VARIABLE CONDENSER—G. W. HAYDEN, Oak Park, Ill. Filed Apr. 16, 1926, issued Mar. 22, 1927.

1,621,722—ELECTROSTATIC CONDENSER—G. W. HAYDEN, Oak Park, Ill. Filed July 6, 1926, issued Mar. 21, 1927.

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1,621,901—ELECTRICAL APPARATUS—W. H. PRIESS, of Montclair, N. J. Filed Dec. 5, 1921, issued Mar. 22, 1927. Assigned to Wireless Specialty Apparatus Company.

1,621,992—VALVE GENERATOR ARRANGEMENT—A. MEISSNER, of Berlin, Germany. Filed Sept. 3, 1921, issued March 22, 1927. Assigned to Gesellschaft Fur Draftlose Telegraphie M. B. H. Hallesches.

1,621,998—ARRANGEMENT FOR ELIMINATING DISTURBING WAVES IN RADIO FREQUENCY CIRCUITS—M. OSNOS, of Berlin, Germany. Filed Apr. 30, 1923, issued Mar. 22, 1927. Assigned to Gesellschaft Fur Draftlose Telegraphie M. B. H. Hallesches Ufer 12/13.

1,622,000—RADIO TELEGRAPH RECEIVING SYSTEM—H. J. ROUND, London, England. Filed Feb. 2, 1923, issued Mar. 22, 1927. Assigned to Radio Corp. of America.

1,622,025—SYSTEM OF RADIOTELEGRAPHY—H. CHIREIX, Paris, France. Filed July 20, 1923, issued Mar. 22, 1927.

1,622,033—RADIOTELEPHONY—A. N. GOLDSMITH and J. WEINBERGER, New York City. Filed May 26, 1922, issued Mar. 22, 1927. Assigned to Radio Corp. of America.

1,622,052—ELECTRICAL CONDENSER—W. J. RICKETTS, Crofton Park, England. Filed Sept. 7, 1926, issued Mar. 22, 1927.

1,622,170—SUPPLY CIRCUITS FOR RADIO SETS—J. G. ACEVES, New York City. Filed Apr. 29, 1922, issued Mar. 22, 1927. Assigned to Crocker-Wheeler Company.

1,622,297—RADIO PRINTING TELEGRAPH SYSTEM—G. S. VERNAM of River Edge, N. J. Filed Dec. 29, 1924, issued Mar. 29, 1927. Assigned to American Telegraph & Telephone Co.

1,622,370—MULTIPLE ANTENNA SYSTEM FOR RADIOCOMMUNICATION—L. ESPENSCHIED, of Queens, N. Y. Filed Sept. 3, 1921, issued Mar. 29, 1927. Assigned to American Telephone & Telegraph Co.

1,622,389—RADIO COMMUNICATION DEVICE—B. F. MEISSNER, of Chicago, Ill. Filed Nov. 6, 1922, issued Mar. 29, 1927.

1,622,459—RADIO DETECTOR—B. H. LUNDQUIST, of Providence, R. I. Filed Feb. 29, 1924, issued Mar. 29, 1927.

1,622,604—BERYLLIUM FILAMENT—A. NYMAN, of New York, N. Y. Filed Nov. 7, 1924, issued Mar. 29, 1927. Assigned to Dubilier Condenser Corporation.

1,622,654—RADIO TRANSMISSION SYSTEM—D. G. LITTLE, of Wilkinsburg, Pa. Filed July 20, 1922, issued Mar. 29, 1927. Assigned to Westinghouse Electric & Mfg. Co.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED MAY 4, 1927

Transferred to Member Grade

Connecticut,	Hartford, 1711 Park St.,	Beekley, F. C.
Pennsylvania,	Pittsburgh, 218 Beech St., Edgewood	Little, Donald G.
England,	Essex, Ilford, 1 Montreal Road,	Cooper, J. A. J.

Elected to Member Grade

Spain,	Cartagena, Submarine Naval Base,	Matres, Trinidad

Elected to Associate Grade

Alabama,	Gadsden, 246 S. 5th St.,	McCormack, J. E.
California,	Berkeley, 2531 Channing Way,	Culley, Lester D.
	Jackson, National Hotel,	van Thiel, Jan
	Los Angeles, 2200 S. Union Ave.,	Hankins, Marvin J.
	San Diego, 4177 Alabama St.,	Doig, John H.
Colorado,	Boulder, R. No. 1, Box 77,	Sammons, Floyd
Connecticut,	South Manchester, Manchester Hospital,	Watt, Earl W.
Delaware	Wilmington, 211 S. Grant Ave.,	Mertz, John M.
Dist. Columbia	Washington, 2017 "O" St., N. W.,	Shortland, James W.
Florida,	Jacksonville, 502 Spearing St.,	Ffoulkes, Cecil E.
	Melbourne, Melbourne Elec. Co.,	Christensen, B. J.
	Tampa, 226 Hyde Park Ave.,	Fisher, Carroll S.
	West Palm Beach, P. O. Box 3243,	Leftwich, E. H.
Illinois,	Chicago, 4940 Lowell Ave.,	Kehm, Clarence H.
	Chicago, 1024 N. Calif. Ave.,	Mazur, Robert S.
	Chicago, 5740 Laftin St.,	McConnell, Harold
	Chicago, 1511 Sedgwick St.,	Parisek, Edward E.
	Decatur, 1538 N. Edward St.,	Spies, Mack C.
	Evanston, 732 Hinman Ave.,	Giard, E. A.
	Fort Sheridan, Signal Corps.,	Maki, George J.
	Maywood, 1926 So. 6th Ave.,	Peary, Malcolm J.
	Quincy, 412 N. 4th St.,	Johnson, J. Robert
Indiana,	North Vernon, Platter Cabinet Co.,	Martin, James A.
Iowa,	Burlington, 914 University Place	Jeffrey, Byron B.
	Cedar Rapids, 526 "C" Ave., East,	Miller, Calir R.
	Dubuque, 3100 White St.,	Long, Frank E.
Kansas,	Ft. Leavenworth, Box 506,	Deremer, Chas. W.
	Lawrence, 200 West 12th St.,	Weedfall, Wm. W.
	Wichita, 509 Winne Bldg.,	Crabtree, Clyde A.
	Wichita, 522 S. Main St.,	Riggs, A. V.
Kentucky,	Berea, Berea College, Box 732	Bryant, Earl A.
	Owensboro, 306 Frederica St.,	Field, Robert W.
Louisiana,	New Orleans, Trop. Radio Tel. Co.,	Whitney, Chas. L.
Maine,	Bangor, 66 Kenduskeag Ave.,	Preston, George D.
Maryland,	Baltimore, 14 S. Broadway	Millhollin, Lorren E.
Massachusetts,	Boston, The Shepard Stores, WNAC,	Stone, C. Willis
	East Springfield, Westinghouse E & M	Coulson, Everett J.
	Haverhill, Postal Teleg-Cable Co.,	Taylor, B. H.
	Roxbury, 3 Westminister Terrace,	Canner, William
	Watertown, 12 Appleton Terrace,	Church, Vallette S.
	Waverly, 105 Beech St.,	McManus, Cornelius F.
Michigan,	Detroit, 11543 Linwood Ave.,	Forna, Alfred A.
	2207 Philip Ave., Apt. 3,	Prange, Harry F.
	Flint, 734 Louisa St.,	Morgan, Ernest W.
	Hartford,	Devore, W. D.
Missouri,	Columbia, Stephens College,	Lewis, Earl W.
Nebraska,	Omaha, 1206 Dominion St.,	Kotera, William J.
	Omaha, Brandies Store,	Thompson, Harry E
Nevada,	Reno, 1035 Sierra St.,	Hauschild, John P.
New Jersey,	Camden, Hotel Camden,	Martin, C. Oakley
	Ventor City, 11 S. Fredericksburg Ave.,	Sharkey, Wm. J.

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New York,	Albion, 17 Lincoln St.,	Yahnke, Wm. J.
	Brooklyn, 2245 East 23rd St.,	Domes, Morris M.
	Brooklyn, 1084 New York Ave.,	Hoefler, Raymond L.
	Brooklyn, 190 Jerome St.,	Kent, Stephen F.
	Brooklyn, 446 Prospect Ave.,	Maxfield, Lewis S.
	New York, 814 Hewitt Place,	Abrahamian, Garo R.
	New York, 522 West 136th St.,	Baldwin, Chester P.
	New York, 2030 Grand Central Ter.,	Bryan, Herman W.
	New York, 1749 Grand Concourse,	Fischer, Leonard O.
	New York, 100 East 42nd St.,	Hunt, Henry T.
	New York, c-o Postmaster, USS Porter	Kortes, George T.
	New York, 345 East 33rd St.,	Mackay, William A.
	New York, 81 East 119th St.,	Nicolai, Erwin W.
	New York, 331 West 83rd St.,	Shine, L. J.
	New York, 4068 Broadway,	Waas, Albert
	Rochester, 15 Caledonia Ave.,	Dodson, George W.
	Schenectady, 103 Nott Terrace,	Brackett, Richard T.
	Schenectady, 1411 State St.,	Dunning, Orville M.
	Stapleton, Staten Island, uscg Colfax,	Hines, Joseph
N. Carolina,	Winston-Salem, Box 806,	Kickler, E. M.
Ohio,	Alliance, 112 S. Liberty Ave.,	Morgan, O. K.
	Cleveland, 3987 West 22nd St.,	Ermatinge, Alex E.
	Cleveland, 3545 Storer Ave.,	Fouse, Harold
	Cleveland, 1071 E. 98th St.,	Hoefele, H. E.
	Cleveland, 1726 St. Clair Ave.,	Talkes, Louis N.
	Cleveland, 18112 Kinsman Blvd.,	Thomas, William W.
	Cleveland, 14709 Colt Road,	White, Louis A.
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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 15 JULY, 1927 Number 7

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GENERAL INFORMATION

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INSTITUTE ACTIVITIES

JUNE MEETING OF THE BOARD OF DIRECTORS

At the meeting of the Board of Direction of the Institute, held at 4:00 P. M. on June 1, 1927 in the Institute Offices, the following were present: Dr. Ralph Bown, President; Frank Conrad, Vice-President; Dr. J. H. Dellinger, Donald McNicol, Melville Eastham, R. A. Heising, R. H. Marriott, L. E. Whittemore, and J. M. Clayton, Assistant Secretary.

Upon recommendation of the Committee on Admissions, the following applications for admission and transfer were approved: Transfer to the grade of Fellow: Dr. A. Hund. Transfer to the grade of Member: E. A. Laport and L. Spencer. Election to the grade of Member: F. K. Dalton and C. L. Lyons.

One hundred and forty-four Associates and nineteen Juniors were elected.

A petition from Institute members residing within the vicinity of Atlanta, Georgia, for the formation of an Atlanta Section of the Institute was approved, and the Section was recognized.

BINDERS FOR COPIES OF THE PROCEEDINGS

From the office of the Institute there are now available binders which hold twelve copies of the PROCEEDINGS. These binders are fitted with wire straps which hold each copy in place. They are finished with Spanish grained and buffed blue fabrikoid with suitable engraved titles on the backbone and front cover. They can be purchased by members of the Institute for $1.50 each.

NEW YORK MEETING

At the meeting of the Institute held on June 1, 1927 in the Engineering Societies Building, 33 West 39th Street, New York City, a paper entitled, "Telephone Communication Over Power Lines By High Frequency Currents" was

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presented by C. A. Boddie of the Radio Engineering Department of the Westinghouse Electric and Manufacturing Company.

General discussion followed the presentation of this paper.

The attendance at this meeting was one hundred and seventy-five.

Copies of the paper, in pamphlet form, are available to members of the Institute upon application to the Institute Offices.

News of the Sections

Atlanta Section

The Atlanta Section held its first formal meeting on the evening of July 6, 1927.

This Section is an outgrowth of the Atlanta Association of Radio Engineers which was formed over three years ago for the purpose of creating sufficient interest, in Atlanta and its vicinity, in radio engineering meetings to enable it to function as a Section of the Institute.

Major Walter Van Nostrand is the Chairman of the Section Organizing Committee.

The Publicity Committee of the Section has been appointed and is composed of the following members: Lambdin Kay, Henry L. Reid and Will Smith.

Buffalo Temporary Organization

A petition for the formation of a Buffalo Section of the Institute has been received at Institute Headquarters.

The officers of the temporary organization are as follows: L. C. F. Horle, Chairman; Dr. L. G. Hector, Vice-Chairman; C. J. Porter, Secretary-Treasurer. Two temporary Committees have been appointed as follows: J. Eichman, Chairman, Membership Committee; I. R. Lounsberry, Chairman, Meetings and Papers Committee.

A meeting was held on June 6th at which time a talk was delivered by Dr. L. G. Hector on "Radio Transmission Phenomena."

The attendance at this meeting was fifty.

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CHICAGO SECTION

A meeting of the Chicago Section was held in the rooms of the Western Society of Engineers on May 31, 1927.

Professor G. M. Wilcox was the presiding officer. A paper was presented by C. W. Horn, of the Westinghouse Electric and Manufacturing Company on the subject of "The Synchronizing of Radio Stations."

CLEVELAND SECTION

On May 27, 1927 a meeting of the Cleveland Section was held in the Case School of Applied Science.

John R. Martin presided. W. C. Blackburn presented a paper on "High Quality Reproduction" with which were included demonstrations.

The attendance at this meeting was fifty-one.

CONNECTICUT VALLEY SECTION

A meeting of the Connecticut Valley Section was held on June 10, 1927 in the Scott Laboratory of Physics, Wesleyan University, Middletown, Connecticut. Dr. W. G. Cady was the presiding officer.

Dr. A. Hoyt Taylor of the Naval Research Laboratory presented a paper on "Some Peculiarities in High Frequency Transmission and Reception."

Messrs. Bourne, Cady, Laport, Warner, Lyford, Kruse and others participated in the discussion which followed.

After adjournment the forty-one members attending the meeting inspected the Laboratory and the research work in progress.

DETROIT SECTION

On April 15, 1927 the Detroit Section held a meeting in the Conference Room, Detroit News Building, 615 West Lafayette Street, Detroit.

Thomas E. Clark presided. A paper by James E. McNary entitled "Some Experimental Radio Field Intensity Measurements and Observations" was read by Mr. McNary.

Twenty-eight members and guests attended this meeting.

The Detroit Section held a meeting on May 20, 1927 in the Conference Room of the Detroit News.

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A talk by Otis H. Trowbridge on "Receiving Set Design" was given. The discussion was participated in by Messrs. Glatzel, Morse, Buchanan, Hoffman and others.

The attendance was thirty-one.

A meeting of the Detroit Section will be held in the Detroit News Building on June 24, 1927, at which time E. D. Glatzel will read a paper, "Some Experiences in Short Wave Transmission."

Los Angeles Section

The Los Angeles Section held a meeting on May 16, 1927 in the Los Angeles Elks Club. L. Taufenbach presided.

Two papers were presented at this meeting. The first, by Mr. Hutson of the Radio Power Corporation of America, was entitled "A-B-C Power Devices, Their Design and Uses." This paper was discussed by Messrs. Wallace, Watters and Chester.

The second paper by W. W. Lindsay, Jr. was on "A Method of Maintaining Constant Radio Frequency Amplification over the Broadcast Range."

The attendance at this meeting was forty.

Philadelphia Section

The Philadelphia Section held a meeting on May 26, 1927 in the Bartol Laboratories, Philadelphia. J. C. Van Horn was the presiding officer.

W. J. Healy deliverey a talk on "Sensitive Measuring Instruments used in Radio."

The attendance at this meeting was thirty-five.

The next meeting of the Philadelphia Section will be held on June 24, 1927 in the Franklin Institute, at which time C. Francis Jenkins will present a paper on, "Radio Vision."

Washington Section

On May 12, 1927 a meeting of the Washington Section was held in Harvey's Restaurant, 11th and Pennsylvania Avenue, N. W., Washington, D. C. Dr. A. Hoyt Taylor was the presiding officer.

Commander S. C. Hooper, U. S. N., delivered a paper,

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"The Mission and Scope of Naval Radio." This paper was discussed by Admiral Bullard and others.

The attendance was forty at this meeting.

Committee Work

COMMITTEE ON SECTIONS

A meeting of the Committee on Sections was held on May 27, 1927. The following attended this meeting: D. H. Gage, Chairman, M. Berger, E. R. Shute and J. M. Clayton.

The Committee reports satisfactory progress on the preliminary draft of a booklet being prepared for the information of Institute members interested in Section formation.

The Committee considered the application from the Atlanta, Georgia members for the formation of an Atlanta Section, and approved the petition submitted.

It is planned to hold additional meetings of this Committee throughout the summer.

COMMITTEE ON MEMBERSHIP

A meeting of the Committee on Membership was held on June 13, 1927.

The Committee has outlined a very intensive plan whereby many desirable and eligible non-members of the Institute who are not familiar with the Institute will receive information regarding it.

Plans for circularization of a number of organizations are being laid with the view of securing desirable members.

TECHNICAL COMMITTEE ON VACUUM TUBES, A. E. S. C.

SECTIONAL COMMITTEE ON RADIO

The Technical Committee on Vacuum Tubes had a meeting on May 4, 1927 at the Institute office. Those present were: A. M. Trogner (acting Chairman), A. F. Van Dyck, C. H. Sharp and A. V. Loughren.

The Committee took action on a number of subjects which have been considered by correspondence in recent months.

The Committee adopted as standard for recommendation to the Sectional Committee on Radio, the two UX bases widely used commercially for vacuum tubes, and also the

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vacuum tube terminology as published in the 1926 Standardization Report of the Institute of Radio Engineers.

There was a discussion of filament voltage tolerances but it was decided that further commercial standardization is necessary before action on this subject would be appropriate.

The Committee took cognizance of a proposal which has come from the International Electrotechnical Commission for standardization of tube bases. The Committee determined to assemble data on the several types of tube bases used in the United States and to place this information at the disposal of Dr. J. H. Dellinger, who will be the United States representative on this subject at the meeting of the International Electrotechnical Committee in Italy in September.

Preliminary Reports of I. R. E. Standardization Sub-committees

As mentioned in this portion of the PROCEEDINGS last month, copies of preliminary drafts of reports of three of the Sub-committees of the I. R. E. Standardization Committee are available for distribution to all persons interested in radio standardization.

Copies of these drafts will be mailed upon application to Institute headquarters.

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TELEPHONE COMMUNICATION OVER POWER LINES BY HIGH FREQUENCY CURRENTS *

C. A. BODDIE

(Radio Engineering Department, Westinghouse Electric and Manufacturing Company.)

High frequency currents have been in use for telephone communication over high voltage power lines for four or five years. In this time a considerable amount of equipment has been installed and is now in operation. The primary object aimed at is to provide a reliable channel of communication between the load dispatcher of a power system and the operators in charge of the various sub-stations under his control.

Before the development of this type of high frequency communication the only system available was the ordinary wire telephone. Power companies were accustomed to provide themselves with communication facilities of their own design and construction, or they might lease circuits from the commercial telephone companies. Companies operating large city systems often found it convenient to lease their telephone circuits from the telephone company. Companies operating long transmission lines however, found that leasing circuits from the telephone company was less satisfactory as well as very costly. In many cases no telephone circuits were available between the points where communication was desired so it became necessary for the power companies to build their own telephone lines. Communication between the load dispatcher and his various sub-stations must be prompt and sure. Much valuable time was lost in going through telephone company exchanges. For this reason privately owned lines were much more satisfactory, due to the promptness with which communication could be established.

It was natural that in building its own telephone lines a power company should follow the same right of way as the

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power line itself, or even to string the telephone wires on the same poles or towers as the power line. Privately owned telephone lines were therefore exposed directly to the induction from the power line for long distances, and were thus likely to be noisy. This was especially the case when the telephone wires were strung on the same poles as the power line.

Excepting for the noise, the privately owned lines usually worked out best. In case of disturbance on the power line such as failure of an insulator, telephone communication was almost invariably interrupted due to blowing of fuses and a discharge over the lightning arresters. This usually occurred on leased lines as well as on the power companies own line. In case of sleet or wind storms the power companies telephone line was often of stronger construction, and was likely to stay up as well or better than telephone companies line. Thus from the viewpoint of continuity of telephone service, leased lines had little to offer as compared with private lines and private lines had a marked advantage from an operating point of view.

Reliable telephone communication is a matter of vital importance to a power company. Its wire telephone circuits being directly exposed to the power line are rather sensitive to disturbances in the power line. The problem of protection is also a rather serious matter when the telephone line is run on the same poles as the power line. This problem has been worked out by the power companies themselves with little or no assistance.

With the rapid development of radio as a means of communication, it was natural that the large manufacturers of power apparatus who had also become deeply interested in radio, should turn their attention to the possibility of using radio apparatus and radio methods in an attempt to provide a better solution for the communication problem of power companies. This work was undertaken by the manufacturer of power apparatus in the belief that anything which assists the power companies more effectively or more efficiently to operate their apparatus, must be a benefit to the whole industry.

The special advantage to be derived by the use of high frequency currents flowing in the power conductors is that the communication circuit is then as strong mechanically as

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the power line itself. The general experience is that wind and sleet which will completely carry away a telephone line, does no great harm to a power line because of its sturdy mechanical construction. The elimination of noise arising from induction from currents in the power line was a second important consideration. The omission of the telephone conductors themselves together with their maintenance and the location of communication equipment at accessible points in power stations and substations, were additional advantages to be gained by the use of high frequency currents.

There are at present three manufacturers of high frequency power line telephone equipment. The equipment of each manufacturer differs rather sharply both in fundamental principles of design and in many minor features from the equipment of the other manufacturers. These striking differences are largely the result of a difference in viewpoint and experience of the engineers making the designs. In addition to a difference in the engineering viewpoint there is also a difference in the theoretical conception, and the analysis of some of the factors entering the problem.

The major outstanding differences in the several designs are:

1. Power level i. e. the amount of high frequency energy which should be provided for a given communication.

2. Number of frequencies used per channel.

3. Method of obtaining duplex in two frequency systems i. e. filters vs. balancing by means of potentials derived from transmitter.

4. Method of calling i. e. interrupted continuous wave (cw), or modulated cw.

5. Type of selector i. e. automatic telephone versus train dispatching selector and code key.

6. Protective equipment.

7. Method of coupling to power line i. e. by condenser or antenna.

Power Level

Perhaps the first and most striking difference in the designs of the several manufacturers is that due to a difference in power level. Two manufacturers have developed equip-

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ment which might be called low power equipment, while one manufacturer provides a transmitter developing so much greater power that it might be called high power by contrast. The low powered equipment provides a transmitting energy of the order of 1 watt, whereas the high powered equipment provides 250 watts. This great difference in power is due directly to the different estimate the various designers place on the value of telephone communication to a power company. The greater the amount of power provided, the greater the margin over the minimum necessary for successful operation. That is, the more power provided for a given communication the greater the factor of safety.

The low powered set is more in line with previous practice in telephone communication over wires. It is also in accord with wire telephone practice which assumes that another circuit is always available in the case of failure of the one under consideration, and that therefore large margins of operating energy are unnecessary. In this case the failure of a circuit means merely switching to another circuit. In this respect a power line telephone is not comparable with the usual wire line telephone, since the failure of the circuit in question means total interruption of communication and this communication is vastly more important than that usually going forward on ordinary commercial telephone lines.

In addition to the much greater importance of load dispatching telephone communication, there are other very strong reasons for using considerable power when the communication circuit is the power line itself. The circuit offered by a power line differs from an ordinary telephone circuit in that it consists not merely in two simple wires connecting the transmitter and receiver, but contains at least three wires and very often six wires. On this circuit is connected a great variety of apparatus. The circuits branch repeatedly and these branches often are tied back into the main circuit forming loops so that the whole becomes a network, rather than simply a line. The transmission properties of this network are not constant, but change sometimes seriously every time a switch is thrown on the high tension system. It is almost impossible to foretell just what will happen when any given switch is opened or closed. In order to provide for these uncertainties, provision must

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be made for an ample supply of communication energy.

Perhaps the most serious difference between a simple telephone line and a communication circuit formed by a power line is that the high voltage gives rise to line noises. All lines are noisy more or less. The magnitude of these line noises vary greatly as between lines. It also varies widely from day to day and from hour to hour. The origin of these line noises seems to be in what are called "spitting insulators". This is a slight static discharge often invisible, occurring between parts of the hardware and the insulator surfaces, or between sections of the insulator itself. This discharge releases the stored energy in the parts involved and constitutes a miniature radio transmitter, which sends a shock disturbance or wave down the line. In character it manifests all the properties of ordinary radio static. The only answer to the noise problem is to increase the communication energy level to such a value that the noise is relatively small in comparison. No amount of amplification at the received end can correct for lack of communication energy.

There has been an attempt to meet the varying conditions of noise and attenuation by providing the transmitter with two sets of tubes, one for low power operation and one for high power operation. As an answer to the line noise problem this is not satisfactory, since with such a system the receiving equipment is always adjusted to a rather high state of sensitivity in order to operate on low power. In case the noise level of the line rises to such a point as to make the use of the high power necessary, the detector is already blocked with noise and when the high power is applied the detector is abused still further. The result is that when the noise level rises to a point which puts the low powered channel out of business the distortion is so great that communication is almost impossible.

From the engineering viewpoint it is clear that it would be unwise to hazard the operation of so important a function as load dispatching telephone communication by the use of low power in an effort to economize. The apparatus discussed in this paper is of the high powered type. In its design it was assumed that the service for which it was intended was of vital importance. It was intended that the reliability and continuity of service should be of the same

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high order as the factor of continuity of the power line itself. In the selection of circuits and the design of the parts the same ample margins are provided over and above the bare operating requirements as are provided in the matter of power. No circuit elements are employed about whose stability and certainty there can be any doubt.

Number of Frequencies Per Channel

The equipment now available for communication over power lines may be divided into two classes:

1. The system providing two frequencies for each channel.
2. The system providing only one frequency per channel.

The two frequency system is the pioneer and still represents the safest and best method of providing duplex communication. In this system each station is assigned two frequencies, one frequency for the transmitter and one frequency for the receiver. The selection of two entirely different frequencies for transmission and reception renders these two functions entirely independent. When properly worked out they are just as separate as if separate wires were provided for each function. With this system the functions of transmission and reception can go forward simultaneously just as in ordinary conversation. The system is very flexible and very stable. It may be modified to do many things which the original design did not contemplate owing to its inherent stability. For example, it may be used as a repeater owing to the fact that the functions of transmission and reception may go forward simultaneously and not alternately.

The single frequency system has been developed along the lines of space radio communication where the transmitter and receiver are tuned to the same frequency, but are alternately connected to the antenna. With this system communication is normally possible in only one direction at a time, it being necessary for each of the parties to throw over in order to reverse the direction of transmission. The first single frequency system was of this throw over type and was manually operated. The superiority of the two frequency duplex system soon drove out the manually operated single frequency simplex system. One company still

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manufactures single frequency simplex equipment, but this manual throw-over has been modified so that it is now automatic. In this system the receivers are normally connected to the antenna in a condition to receive and the transmitters are inactive. A voice operated relay is used to unlock the transmitter and allow it to oscillate and at the same time block the receiver, thus accomplishing automatically the throw-over which was formeerly manual. The system is

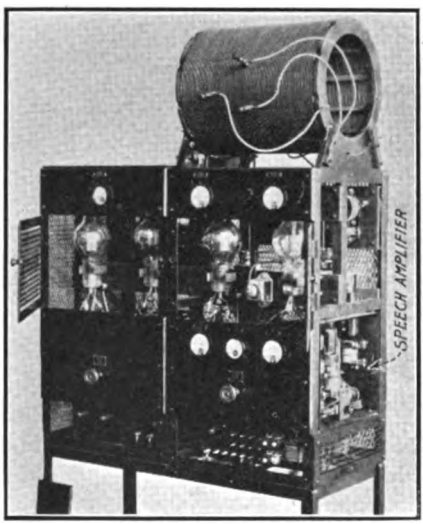

Fig. 1.

not really duplex, but is automatic simplex. The functions of transmission and reception go forward alternately and not simultaneously as is essential in true duplex. It is not so flexible as the two frequency system nor so stable. It cannot be used as a repeater without complete duplication of equipment, owing to the fact that the functions of transmission and reception take place alternately, whereas a repeater requires them to go forward simultaneously. Such a system is much more susceptible to noise picked up on a line extension. If the noise level is appreciable it will seriously interfere with the operation of the equipment.

It is believed that the two frequency duplex system will

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be finally adopted owing to its superior stability and flexibility.

General Description of High Powered Set

A view of the transmitter of the high powered equipment is shown in Fig. 1. It is built around a 250-watt tube as the fundamental element. The tubes used have plain tungsten filaments and were selected because of their stur-

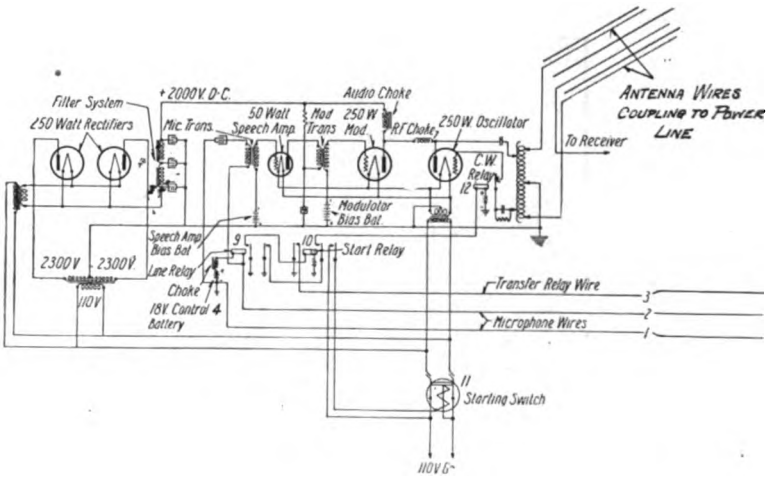

Fig. 2.

diness and great reliability. The transmitter has two power tubes, one used as an oscillator or power amplifier and one used as a modulator. A 50-watt tube is used as a speech amplifier to control the grid of the 250-watt modulator. The speech amplifier is seen at the right of Fig. 1. The power supply is obtained directly from the 110-volt, 60-cycle system or from a motor generator converting from d-c. to a-c. if the 60-cycle supply is interrupted. This voltage is stepped up by means of a transformer and rectified in the usual manner to provide 2000 volt d-c. plate supply for the transmitter. The rectifier is seen at the left of Fig. 1. The general nature of the circuits used is shown schematically in Fig. 2. These circuits are those commonly used and need no further description. In order to reduce the number of wires on the diagram, the customary method of showing a separate battery and ground for each relay is employed.

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A general view of the receiver used is shown in Figs. 3 and 4. The circuits employed are shown in Fig. 5. The apparatus is grouped and mounted on separate panels. The top panel Fig. 3 contains all the radio circuits. The second panel contains all vacuum tubes, transformers and accessories which have to do with detection and amplification. The third panel carries the plate rectifier for supplying plate voltage to the amplifier tubes. It also carries a group of B

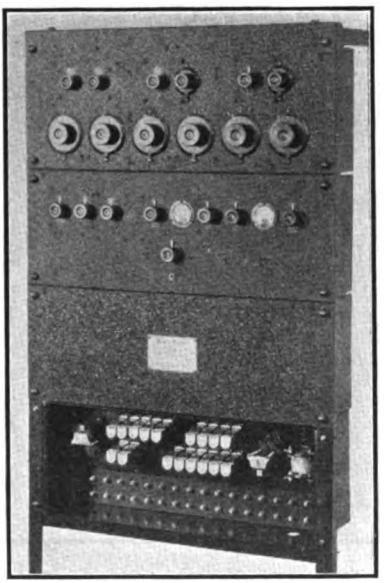

Fig. 3.

batteries which float on the plate supply rectifier for the purpose of stabilizing the voltage. This B battery does not furnish plate current except in emergencies when the power supply is off or the voltage low. The lower panel carries all control relays and provides an ample terminal board.

The system is entirely automatic. It is controlled by a standard form of automatic telephone desk stand. When the receiver is lifted from the hook, current is drawn from an 18 volt control battery shown in Fig. 2 over the microphone wires 1 and 2. This current is drawn through relay 9 which picks up and energizes relay 10, which in turn closes the starting contactor 11, and applies 110 volts, 60 cycles to

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the transmitter. The tubes light and the plate rectifier supplies the necessary 2000 volt d-c. potential to the transmitter. The output of the transmitter is fed into the power line by means of coupling wires or antenna associated with the power line.

Calling is accomplished by interrupting the flow of high frequency current by means of the calling dial on the desk

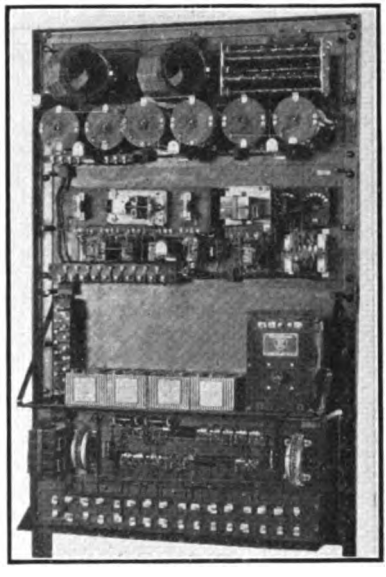

Fig. 4.

stand. When the calling dial is operated, the microphone circuit is momentarily interrupted, which causes relay 9 to drop out at each interruption. A circuit is formed through a back contact on one of the armatures of relay 9 and a front contact on relay 10 to energize relay 12. This relay opens the grid circuit of the oscillator and thus stops the flow of high frequency current in the power line. Thus at each interruption of the microphone circuit by the calling dial there is a corresponding interruption of the flow of high frequency current into the power line. The number of these interruptions is controlled by the number on the calling dial.

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Heterodyne System of Calling

The circuits of the receiver are of the simple coupled circuit type. The tubes are kept lighted at all times and the detector is in a state of continuous oscillation at all times except during talking. When a signal is received a beat note is produced by the oscillating detector in the usual manner. The amplified beat note is fed into a pair of tubes in parallel indicated at 15, Fig. 5. These tubes are provided with grid leak and condenser and act as detectors. The plate current

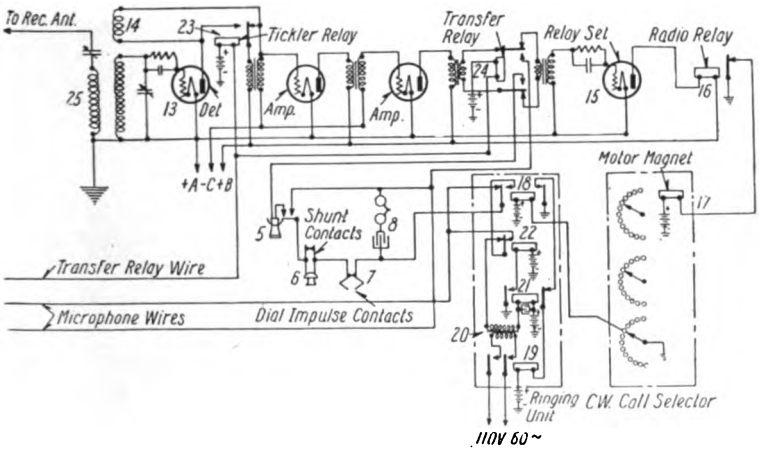

Fig. 5.

of these tubes is drawn through a polarized relay 16 which is normally energized. The incoming signal reduces the plate current and allows the relay armature to drop out and close the contacts. Two tubes are used in order to insure ample power to actuate the relay contacts. The heterodyne system was adopted because of its well known superiority over all other methods of reception. It has demonstrated its value in the service of power line communication. With this system it is possible to call through disturbances and line conditions which it is impossible to talk through. This is just the reverse of the situation when the earlier system using modulated impulses was used. It has a singular ability to work through line disturbances which would put other methods previously used out of business.

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SELECTOR

The selector used is one of the simple types used in automatic telephony. It consists of a motor magnet shown in Fig. 6, which by means of a ratchet and pawl drives an arm over a bank of contacts. At each impulse of the calling dial the motor magnet is advanced one step. By means of a combination of fast and slow relays a circuit is closed at the final pause in the dialing operation, which rings a bell in

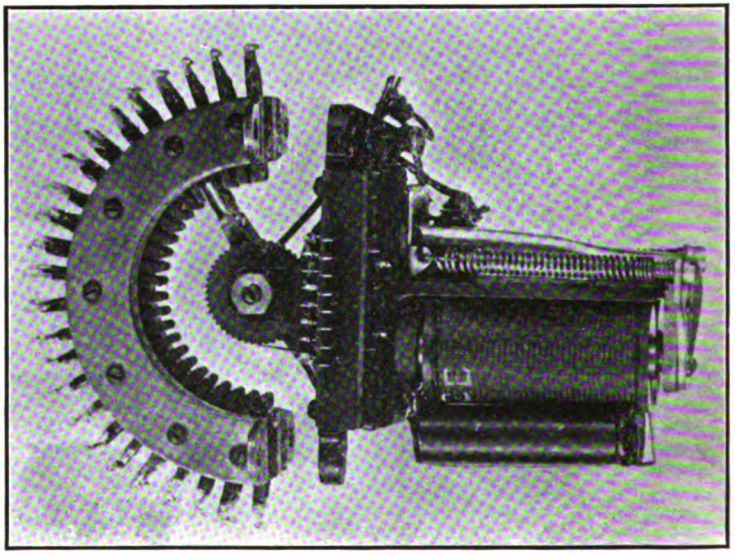

Fig. 6.

case the number dialed corresponds to the number of the station called.

When the party called answers, by lifting the receiver, a relay in the receiver is energized which stops the oscillating detector. Another relay transfers the output of the amplifier from the relay tubes to the desk telephone receiver. The two parties are now in two way communication just as in the case of an ordinary wire telephone line.

MODERATE POWERED SET

There are many applications for communication over power lines where the voltage is low and the distance is

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short. For this class of service a moderate powered set has been developed as shown in Fig. 7. It is entirely self-contained, the one unit carrying the complete transmitter, receiver and selective equipment. The circuits are substantially the same as on the large set. The transmitter consists of five tubes, two of which are power amplifiers and two modulators, one being used as a master oscillator. A new type coated filament tube is used which gives the set a rat-

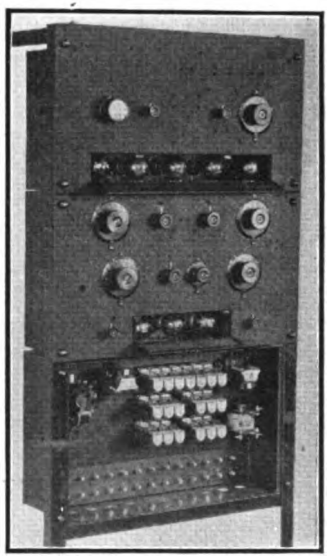

Fig. 7.

ing of 30 watts. This little set is much more powerful than the low powered sets now in use. The receiver consists of three tubes, one-step radio, a detector, and one step of audio amplification.

Protective Apparatus

The protective apparatus required for this class of service requires special consideration. The equipment used is shown in Fig. 8. It comprises a drain coil, two spark gaps, a high voltage fuse and a protective condenser. The elements are connected as shown in the diagram. It is not intended to attempt to open the circuit in case of contact be-

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tween the antenna and a power conductor. The rush of current in this case is so great that nothing less than large station oil circuit breakers could be expected to open the circuit. The intention here is to divert the current rush to ground and clear the apparatus from the main circuit. The spark gaps are of very sturdy construction, having sufficient metal to prevent their being volatilized during the time the station circuit breaker is opening. The gap nearest to the apparatus is set somewhat closer than the one connected directly to the antenna. The drain coil which is of

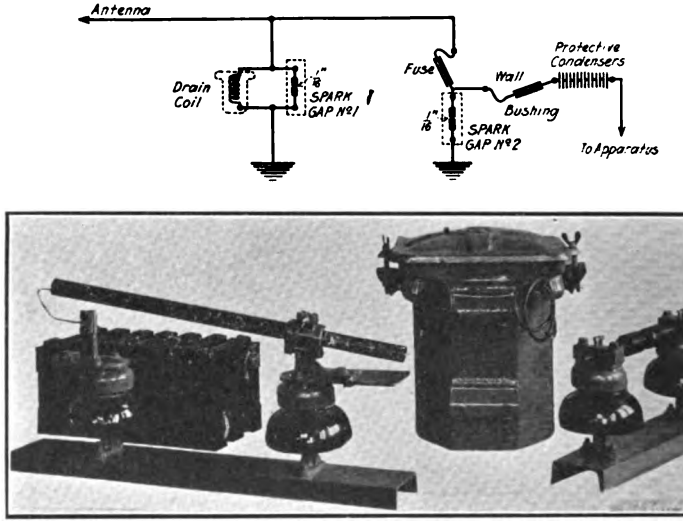

Fig. 8.

relatively fine wire may volatilize and clear itself. If this occurs the main current rush goes through the fuse and the second spark gap. In the act of trying to clear, the fuse will raise the potential sufficiently to flash over the first spark gap, which then takes the discharge and allows the fuse to clear. This was the first adequate protective system developed for this kind of service. It has demonstrated its efficiency in the several instances where it has been called upon to act.

Method of Coupling to Power Line

The method of coupling to the power line which should be adopted is one over which there is considerable difference

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of opinion. Two methods of coupling the communication equipment to the power line are now in use.

The first method is to run antenna or coupling wires parallel to the power line and to use these wires to induce currents in the line, and to receiver currents from the line. This method is commonly called antenna coupling.

Fig. 9.

The second method is to connect the communication equipment to the proper conductors of the power line by suitable high voltage condensers. This method is called condenser coupling.

It is said and quite generally believed that "condenser coupling is much more efficient than antenna coupling". This statement seems so self-evident that it is generally accepted without question. The basis of the claim is found in the fact,

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that, of the total electric field emanating from an antenna wire, only a small part is intercepted by an intervening power conductor. This means only a small part of the antenna current flows in the power conductor, and it is taken to indicate that the antenna is inefficient. The measure of its inefficiency is thought to be represented by the ratio of

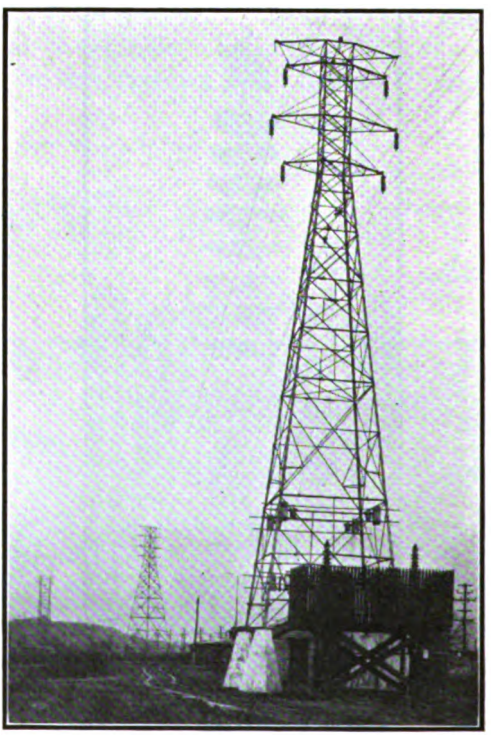

Fig. 10.

the induced current in the power conductor to the inducing current in the antenna.

Early Condenser Installation

If this view were correct, there would be a very great advantage in directly coupling the transmitter to the power conductor by condensers. Acting on this supposition, a set of condensers was built and installed on the lines of the Duquesne Light Company at Brunots Island where they still exist. These condensers are believed to be the first coupling

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condensers to be connected to a high voltage line in this country. Figs. 9 and 10 give a view of this original condenser installation. Two condensers were used, each consisting of a 132,000 volt condenser type bushing supported in a tank filled with oil. This condenser bushing is a standard bushing used in high voltage transformers and switches. It consists of layers of micarta insulation between which are interposed sheets of tin foil. The micarta paper and tin foil are wound on a central brass tube layer on layer.

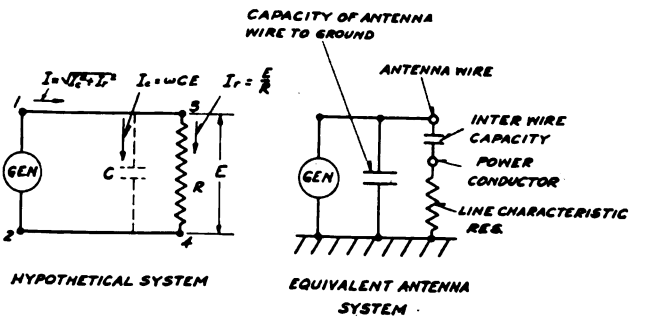

Fig. 11.

The length of the layers of tin foil is reduced progressively as the diameter increases so that the capacity between layers is a constant, the object of the design being to provide a uniform voltage stress on all parts of the insulating material. The capacity of such a bushing is .0003 μf.

Results of Tests

After these condensers were connected to the line, tests were made and compared with the results obtained using antenna coupling. It was found that the signal received at the distant end of the line was not appreciably different from that obtained when using antennae. This was something of a surprise but the reason was soon discovered. The fact is that *antenna coupling is not "inefficient"* as had been previously supposed.

The fact that such a large part of the total electric field from an antenna goes to ground, or to the other antenna wire, does not in itself mean loss, in the sense that there is a loss of energy. The true nature of the relation between condenser coupling and antenna coupling can be obtained

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readily by reference to Fig. 11-a, where a non-inductive resistance is connected to terminals 3, 4 the energy being supplied from a generator at terminals 1, 2. The current through the resistance is determined by ohms law $I_r = \frac{E}{R}$. The energy dissipated in the resistance is $\frac{E^2}{R}$. The generator current under these conditions is I_r as determined above. If now we connect across the terminals 3, 4 a condenser and still maintain the same potential E, the same current will flow through the resistance and the energy dissipation will be $\frac{E^2}{R}$ as before. The current flowing through the condenser will be $I = \omega CE$. The total current carried by the generator is the resultant of these two currents or $I = \sqrt{I_r^2 + I_c^2}$. The generator current is now much larger, but the power factor is correspondingly reduced so that the energy supplied is just the same as before. It is obvious in this case that from the mere fact that the entire generator current does not flow through the work circuit, it does not follow that there is a loss in efficiency.

The above hypothetical case represents exactly the conditions existing when a high frequency transmitter is coupled to a power line by antennae. The useful current delivered to the power line flows against a considerable resistance. If the line is very long this resistance is the characteristic resistance of the line. It amounts to some 800 or 900 ohms for a typical power line at 50,000 cycles as measured between wires or approximately half this value between one wire and ground. If the transmitter is coupled to the line by condensers the entire output is fed directly into the line characteristic resistance. This resistance being high, the current flowing into the line is small. If an antenna is used to couple the transmitter to the line characteristic resistance, there exists a considerable capacity shunted across the work circuit, exactly as indicated in Fig. 11-b. If we consider the case of coupling "phase to ground", the capacity of this shunting condenser is the capacity of the antenna wire to ground minus the capacity from antenna to power conductor.

The current going into the power line with antenna coupling is precisely the same as with condenser coupling,

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if we assume the same amount of energy to be delivered by the transmitter.

The difference between antenna coupling and condenser coupling lies in the existence of a relatively large shunting condenser in the case of antenna coupling and the practical elimination of this shunt capacity in the case of condenser coupling. Both methods provide the same interwire coupling capacity. This shunting capacity does no particular harm since the dielectric is air and therefore the losses are negligible.

Coupling Viewed From Law of Conservation of Energy

The above conclusion may also be arrived at more generally from the view point of the law of conservation of energy. If we assume the shunting condenser "*C*" Fig. 2 has no resistance, it follows immediately that the entire output of the generator must appear in the load resistance 3, 4, Fig. 2, since there is no other place for the energy delivered by the generator to go.

Applying this to the case of the antenna it assumes that the antenna wires themselves have no resistance, and that there is no other loss such as radiation or dielectric absorption in which energy might be dissipated. These two assumptions are not strictly accurate since the antenna wires do have some resistance and there is some radiation, but under the conditions obtaining in practice, these values are so small as not to effect seriously the result.

General Analysis of Antenna Coupling

While the elementary analysis on the basis of conservation of energy is quite sufficient to dispose of the prevalent belief in the inefficiency of antenna as a means of coupling to a power line, the idea is so widespread that it is felt that a more detailed analysis of the mechanism of antenna coupling might be helpful. The mechanism of the electric field and its fundamental theorems have been pretty well established since the days of Faraday and Maxwell. In the development of various phases of the electrical industry practical problems are continually presenting themselves, which require a correct understanding of fundamental relations. The phenomena or mechanism of antenna coupling

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to a power line is one of these. In order to show just how the theorems on the electric field apply to the problem of antenna coupling, it may be desirable to plot the electric field in terms of lines of force and equi-potential surfaces and point out the reasoning on which this is based.

Elementary Principles

In order that the discussion be complete it is necessary to reproduce the equation determining the potential of the point in space due to a charged wire. The electric strains in free space emanating from a charged sphere act in radial lines. The force at any point is measured by the mechanical pull on a unit charge. A unit charge is defined as being such that it will repel an equal charge with a force of one dyne if the distance between the charges is one centimeter. The force at any point is calculated by

$$F=\frac{Q}{r^2} \tag{1}$$

The quantity of electric flux emanating from a charged body is measured in lines or rather in tubes of force. This flux is measured by the mathematical quantity obtained by multiplying the component of electric force resolved perpendicular to the surface by the area over which the force acts. If the force varies from point to point over a surface, this flux is calculated by integrating the normal component of force over the area. This quantity is called "normal induction" and is expressed

$$\Phi=\int NdS \tag{2}$$

where N is the component of force normal to the element of area dS.

The quantity of flux passing through the surface of an imaginary sphere having a charge Q at its center is obtained by multiplying the actual force at the surface of the sphere by the area of the sphere, since the force in this case is everywhere normal to the surface thus

$$\Phi=\int NdS=\frac{Q}{r^2}\times 4\pi r^2=4\pi Q \tag{3}$$

It is to be noted that the quantity of flux is dependent only on the value of the charge within the sphere and is independent of its radius. It is also true that the quantity of flux

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passing through the sphere is exactly the same whether the charge Q is located at its center or at any other point within its interior.

This theorem is not limited to a sphere but is quite general in that the flux passing through any closed surface, regardless of its size or shape is $4\pi Q$. This may be demonstrated by reference to Fig. 12. Let A, B, C represent any

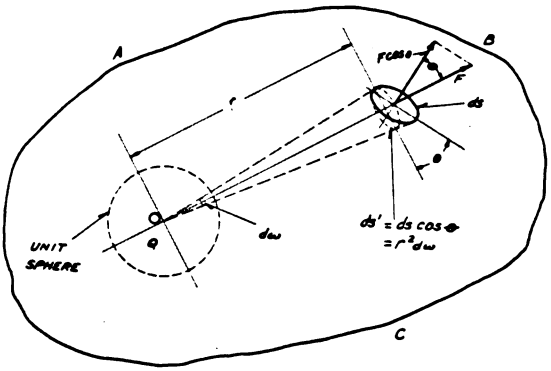

Fig. 12.—Flux through a closed surface $4\pi Q$.

closed surface in which is located a charge Q at some point O. Consider any small element of the surface dS. The force acting at the surface dS is $F=\frac{Q}{r^2}$. In order to calculate the normal induction through dS, we must resolve the force perpendicular to the surface dS. If the surface makes an angle θ with a plane perpendicular to the radius, the value of normal induction for the area dS is

$$N\,dS=F\cos\theta\,dS=\frac{Q}{r^2}\cos\theta\,dS \qquad (4)$$

Consider the cone whose vertex is O and base dS. Let the area subtended at the surface of a unit sphere at O be dw. The area subtended at the surface of a sphere of radius r is $dw.r^2$. This area is equal to the projection of dS along the radius r, that is

$$dS'=dS\cos\theta^2=dw.r^2 \qquad (5)$$

Substituting this in the above equation we get

$$NdS=\frac{Q}{r^2}\times dw.r^2=dw.Q \qquad (6)$$

It is to be observed that the flux passing through the sur-

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face dS is independent of the distance of the charge or the inclination of the surface to the radius. The total flux passing through the surface is found by integration.

$$\Phi = \int NdS = Q\,dw = 4\pi Q \tag{7}$$

the integral of dw being 4π since this is the area of a unit sphere. The normal induction over a surface of any shape due to a charge Q at any point within the surface is $4\pi Q$.

Field Around a Wire

These principles may now be used to calculate the field distribution around a wire in space. Referring to Fig. 13 let A, B represent a wire of diameter d. Describe a cylin-

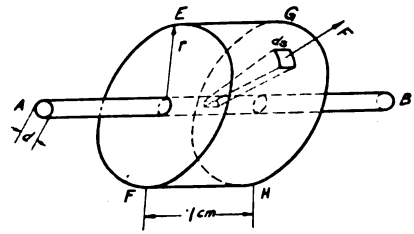

Fig. 13.—Force at any point $= \frac{2Q}{Y}$

drical surface around the wire of radius r. Let EF and GH be two planes perpendicular to the wire and one centimeter apart. Let Q be the charge per unit length on the wire. If F is the force at the surface of the cylinder, the normal induction over the cylindrical surface bounded by the plane is

$$\Phi = \int NdS = F\,(2\pi r) \tag{8}$$

Now from the foregoing theorem the total induction over any surface having in its interior a charge Q is $4\pi Q$. Since by symmetry there is no force parallel to the conductor the entire flux must pass out radially through the cylindrical surface and must therefore be equal to the value calculated above. Hence we may write

$$F(2\pi r) = 4\pi Q. \tag{9}$$

$F = \frac{2Q}{r}$ which gives the force at any distance r.

Electric potential is defined in terms of work. The potential difference between two points is the work required

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to carry unit charge from one point to the other. If we consider a unit charge at a distance r from the center of the conductor to be moved radially outward through a small distance dr, the work done is $F\,dr$. This work is a measure of the change in potential in moving through the distance dr. The potential will fall as we move away from the con-

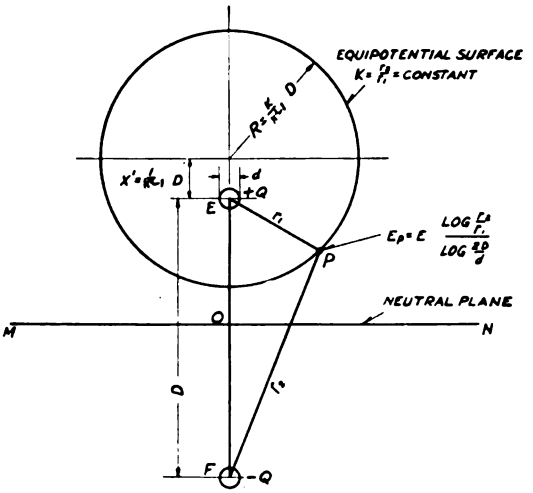

Fig. 14.—Equipotential surface.

ductor, hence the increment in potential is negative and we may write

$$-dE = F\,dr = 2Q \cdot \frac{dr}{r} \qquad (11)$$

integrating this we get

$$-E = 2Q \log r + C \qquad (12)$$

where C is the constant of integration and is to be determined by conditions.

Field Between Two Oppositely Charged Wires

We may now apply this equation to the determination of the field between two wires. Referring to Fig. 14 let P be any point at distances r_1 and r_2 from two cylindrical conductors of diameter d and distance D, having charges $+Q$ and $-Q$ respectively. The potential at the point P is the

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resultant of the charges on the two wires. The potential due to the positive charge is

$$-(2Q \log r_1 + C) \tag{13}$$

The potential due to the charge on the negative wire is

$$+(2Q \log r_2 + C) \tag{14}$$

The resultant potential is the algebraic sum of these which gives

$$E_p = 2Q(\log r_2 - \log r_1) \tag{15}$$

$$= 2Q \log \frac{r_2}{r_1}$$

This equation enables us to calculate the potential at any point P in space provided we know the charge Q per unit length of conductor. It is one of the most important equations having to do with the electric field, due to charges on wires. The factor $2 \log \frac{r_2}{r_1}$ is known from the physical dimensions of the system. This factor is called a potential coefficient since when multiplied by the charge Q it gives the potential of P. The value of Q in any given case is to be determined by the conditions of the problem.

If the potential between the wires is known, we can determine the value of Q in equation 15 by moving the point P toward the wire E until it coincides. It will then have the same potential as the wire and equation 15 becomes

$$E = 2Q \log \frac{D}{\frac{d}{2}} \tag{16}$$

or

$$Q = \frac{E}{2 \log \frac{2D}{d}} \tag{17}$$

Substituting this value of Q in equation 15 we get

$$E_p = \frac{E \log \frac{r_2}{r_1}}{\log \frac{2D}{d}} \tag{18}$$

This is a most important practical working equation since it enables us to calculate the potential of any point in space whose distance r_1 and r_2 from the charged conductors are known.

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Capacity of a Wire to Ground

The capacity of a body or a system is measured by the charge required to raise its potential from zero to unity when all other neighboring bodies are grounded. Equation 17 gives the capacity of a cylindrical conductor with respect to the ground plane since

$$C=\frac{Q}{E}=\frac{1}{2\log_e \frac{2D}{d}} \qquad (19)$$

This gives the capacity in electrostatic units.

When converted into electromagnetic units it becomes

$$C=\frac{0.0388}{\log_{10}\frac{2D}{d}}\ \mu\text{f.} \qquad (20)$$

Equipotential Surfaces

It should be noted that the potential (equation 18) is dependent upon the ratio of the distances and not upon their actual magnitude. Thus if a series of points be so located that the ratio of the distances $\frac{r_2}{r_1}$ is constant these points will all be at the same potential. Such a series of points determine an equipotential surface.

If a curve is drawn so that the ratio of the distances from any point P Fig. 14 to two fixed points E and F is a constant $K=\frac{r_2}{r_1}$ then the curve is a circle. The radius of the circle is $R=\frac{K\,D}{K^2-1}$ and its center lies on the line E, F produced at a distance $X'=\frac{D}{K^2-1}$ above one of the fixed points E. Such a circle is a section of an equipotential surface whose potential is defined by equation 18. A system of such circles representing sections of equipotential surfaces may be drawn for uniform increments of potential and is useful in mapping out the field.

Fig. 15 represents the equipotential surfaces due to two conductors E, F one inch in diameter and 100 feet from conductor to the neutral surface $M\,N$. The curves are drawn at intervals of 5%. The potential between the conductor and the neutral plane is taken as unity. For example, a point 21 feet above the neutral plane $M\,N$ on the line $O\,E$

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has a potential equal to 5% of the potential between the neutral plane and the conductor *E*. A point 10 feet below *E*

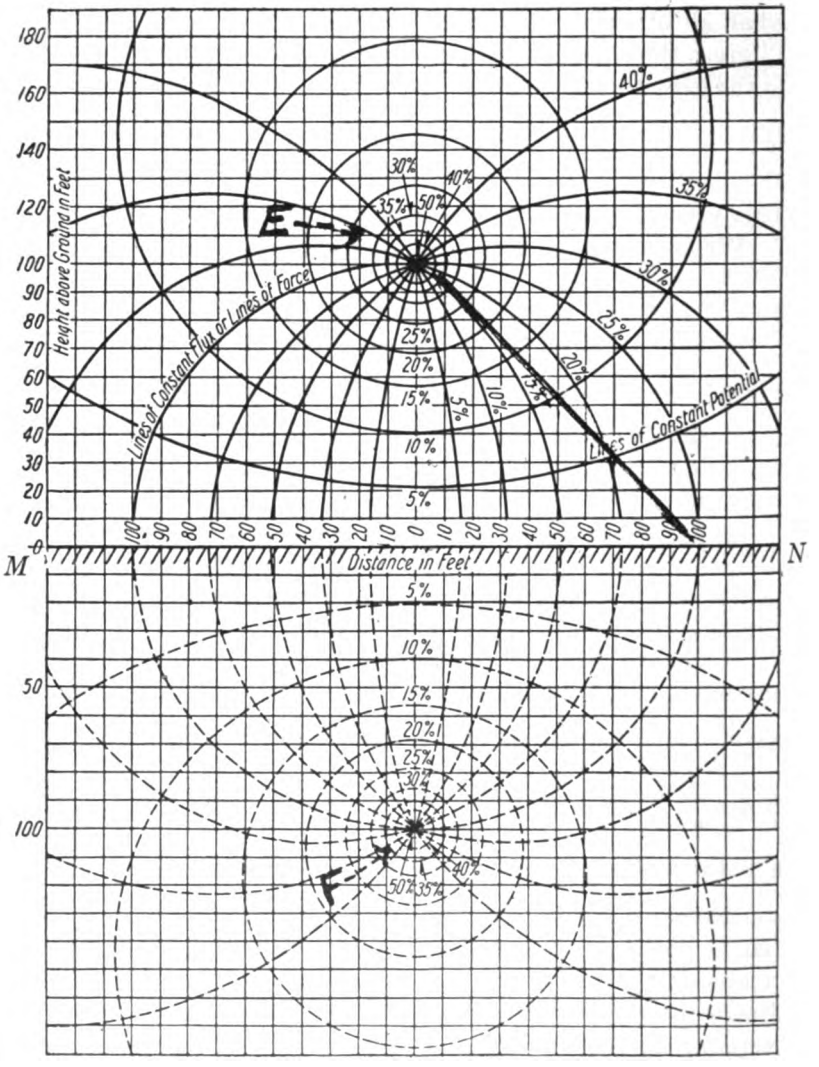

Fig. 15—Electric field due to a wire 1 inch diameter and 100 feet above ground plane.

has a potential 35% of the conductor potential. The potential gradient is greatest nearer the wire, 65% of the total drop occurring in the first 10 feet while the remaining 35%

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is spread out over a distance of 90 feet. Notwithstanding this fact, points at a distance of 20 to 30 feet from the conductor still have a very considerable potential. For example, a point 22 feet below the conductor E has a potential of 25% of the conductor potential and a point 44 feet below has a potential of 15%. As the potentials get lower the centers of the circles are higher and the radius gets larger until at zero potential the circle has an infinite radius and its center is at an infinite distance above E thus giving the straight line $M\ N$. The closer we approach to the charged conductor the higher the potential of the circle and the more nearly its center coincides with the center of the conductor.

The values given on the equipotential circles in Fig. 15 apply not only to a conductor 100 feet above the neutral

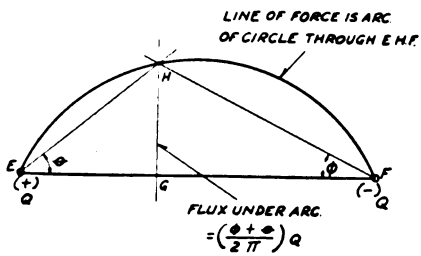

Fig. 16.—Line of Force.

plane and 1 inch in diameter but apply to any conductors so long as the ratio between the conductor diameter and the distance between conductors is constant. For example, the diagram would apply directly to the case of a conductor ½ inch in diameter and 50 feet above the neutral plane, or to a conductor ¼ inch in diameter and 25 feet above the neutral plane, or to a conductor ¼ inch in diameter and 25 feet above the neutral plane. The distances are then not measured in feet but on the basis $O\text{-}E = 100$ and read in % of O E. The scale of the diagram is such that it will apply well to conductors and spacings of power lines.

Lines of Force

The lines of force may also be plotted on the diagram. The method of accomplishing this may be developed by reference to Fig. 16, where E and F represent the two opposite

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charged wires. Considering each conductor by itself, the electric flux will pass out into space radially and uniformly in all directions and the quantity of flux passing through any given area is proportional to the angle subtended at the center of the wire. For example, the flux issuing from conductor E and passing through the plane G, H seen on edge is proportional to the angle θ subtended at E by the line G, H. Similarly the flux passing through G, H due to the conductor F is proportional to the angle θ subtended at F by the line G, H. The total normal induction, that is the total flux, passing through G, H is the arithmetic sum of that due to each conductor individually since they are on opposite sides of the plane G, H and the charges are of opposite sign making the effects additive. The flux passing from E through the angle θ is that part of the flux Q which the angle θ is of the circumference 2π. That is, the flux passing through G, H issuing from E is $\frac{\theta}{2\pi} Q$ and the flux issuing from F is $\frac{\Phi Q}{2\pi}$. The total flux crossing the line G, H is $\frac{\theta + \Phi}{2\pi} Q$. Thus the flux passing between the point H and the line joining the centers of conductors E, F depends only upon the sum of angles θ and Φ. A curve described by the point H, such that the sum of the angles remains constant will enclose a constant amount of flux. The curve so described will be at every point tangent to the resultant electric force and is called a line of force. Every such line of force, between two parallel wires, is the arc of a circle since it is a property of a circle that the angle subtended by a chord at any point on the arc is constant. The line E, F is a chord and the point H is a point on the arc drawn through these three points, E, H, F. The circular arc E, H, F is a line of force passing through H and encloses the total and constant quantity of flux $\frac{\theta + \Phi}{2\pi} Q$.

Referring now to Fig. 15 the lines of force are laid out so as to enclose definite quantities of electric flux. Radial lines are drawn from the point E until they intercept the line M N. A circle is then drawn through the point of intersection H and the two points E, F.

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The percentages marked on the flux curves are the percentages included between the flux curve and the vertical E, F. The value on each curve is $\frac{\theta}{2\pi}$ since the flux intercepted by the line O, H is $\frac{\theta}{2\pi} Q$.

The lines of force and equipotential surfaces are everywhere at right angles.

Neutral Plane Replaced by Ground Plane

If we assume the neutral plane $M\ N$ to be replaced by a conductor and that this conductor is maintained at zero potential the electric field will be everywhere undisturbed. The field above the line $M\ N$ may therefore be taken to represent the field of a conductor E above the ground plane by the height E, O. The conductor F below the ground plane is called the image of E and is useful mainly in deriving the field distribution above the ground plane.

Application of Principles to Antenna Coupling

We may now apply these principles to the problem of antenna coupling. Referring to Fig. 17 let E be the antenna wire and let P represent a power conductor on an equipotential surface E'. If the conductor P is insulated and is of small dimensions relative to the height its presence will not disturb the electric field.

If the conductor P is connected to ground, a charge Q' will flow from ground up into the conductor just sufficient to lower the potential of the conductor P to zero. This quantity of electricity is exactly the same as that necessary to raise the potential of the conductor P to the potential E' Hence

$$Q' = C' E' \tag{21}$$

where C' is the capacity of the conductor P to ground. The reaction of the charge Q' on the charge Q is here neglected for simplicity since its effect is small.

So far we have been considering the question of field distribution in the steady state, that is, we have assumed the potentials and charges to be constant. If we now assume an alternating potential to be applied to the antenna wire,

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the charges and field distribution at any instant may still be determined by the foregoing methods.

If the potential E of the antenna wire varies according to a sine law, the instantaneous potential is

$$e=E \sin \omega t \tag{22}$$

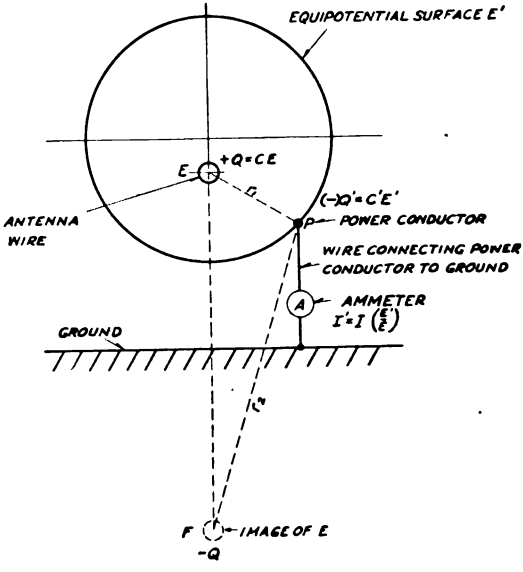

Fig. 17.—Charge induced on grounded power conductor Q' = C'E'.

where $\omega=2\pi f$ and E is the maximum value of e. The value of the induced charge q' at any instant is also a sine function.

$$q'=Q' \sin \omega t. \tag{23}$$

The current passing through an ammeter in the ground wire at any instant is the rate of change of charge or

$$i'=\frac{dq'}{dt}=\omega Q' \cos \omega t \tag{24}$$

The maximum value of the induced current i' is

$$I'=\omega Q' \tag{25}$$

and since $Q'=CE'$ we have

$$I'=\omega C'E' \tag{26}$$

In the same way, the current flowing in the antenna wire is the rate of change of the charge hence

$$I=\omega CE \tag{27}$$

The ratio of the induced current in the grounded power

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conductor to the inducing current in the antenna is

$$\frac{I'}{I} = \frac{\omega C' E'}{\omega C E} = \frac{E'}{E} \tag{28}$$

The capacity C' of the power conductor is in practice very nearly equal to the capacity of the antenna wire and hence cancels out in the above equation.

Equation 28 gives a most important relation in the theory of antenna coupling. It states that the ratio of the current induced in the power conductor to the inducing current in the antenna is the same as the ratio of the potential induced at the location of the power conductor to the inducing potential of the antenna.

The potential ratio $\frac{E'}{E}$ may be determined from the dimensions of the system by equation 18. The induced current may then be expressed as,

$$I' = I \frac{E'}{E} \tag{29}$$

$$= \frac{I \log \frac{r_2}{r_1}}{\log \frac{2D}{d}} \tag{30}$$

Interwire Capacity

The capacity between two wires in free space may be obtained from equation 20 which gives the capacity from one wire to the neutral plane. The mistake is commonly made in supposing that this equation will always give the correct capacity between two wires without regard to other conditions. It should be clearly recognized that this equation is limited specifically to the case where the two wires are in free space and have *equal and opposite charges*. Unless these conditions are fulfilled it does not apply. For instance it does not apply to the case of two wires above a ground plane when one wire is grounded. The field distribution in this case is totally different from that shown in Fig. 15 to which equation 20 applies.

In general the capacity of a body is the charge on the body per unit potential. The capacity between two bodies in a system is measured by the charge or quantity of electric flux leaving one body and terminating on the other, divided by their potential difference, all other bodies being

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held at zero potential. In the case of an antenna wire and grounded power conductor the capacity is measured by the charge on the grounded power conductor divided by the potential of the antenna to ground. The induced charge on the power conductor is given by equation 21 as

$$Q' = C' E'$$

This means that a quantity of electric flux equal to Q' leaves the antenna wire and terminates on the grounded power conductor. The capacity between the antenna wire and the grounded power conductor is then measured by this quantity of flux or charge divided by the potential between the two wires, that is

$$C'' = \frac{\textit{induced charge}}{\textit{potential difference}} = \frac{Q'}{E} = \frac{C' E'}{E} \qquad (31)$$

which may be written

$$C'' = C\left(\frac{E'}{E}\right) \qquad (32)$$

since $C' = C$ very nearly. That is, the capacity effective between the antenna wire and a power conductor is equal to the capacity of the antenna to ground multiplied by the potential of the equipotential surface on which the power conductor is located measured in per cent.

If the power conductor were not present, or if it were insulated from ground, the total electric flux emanating from the antenna would all fall on the ground and the field distribution would be as shown in Fig. 15. When the power conductor is grounded, the total flux issuing from the antenna does not change much, but the distribution is changed considerably. The total antenna flux has then two paths to ground, one by way of the power conductor and the other direct to ground. The part going to ground through the power conductor is merely diverted from its natural path. The part going direct to ground may be obtained by subtracting that through the power conductor from the total. We may obtain the effective antenna capacity to ground by subtracting the interwire capacity which is proportional to the interwire flux from the total antenna capacity which is proportional to the total antenna flux.

$$C_g = C - C'' \qquad (33)$$
$$= C\left(1 - \frac{E'}{E}\right)$$

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Calculation of Antenna Efficiency

These relations may now be used to determine the efficiency of antenna coupling. As an example let us assume the simple case indicated in Fig. 17, where E represents the antenna wire, 50 feet above the ground plane, and P represents the power conductor 10 feet below the antenna. Assume the diameter of both wires to be .5 inches. The potential of the point P with respect to the antenna potential may be calculated from equation 18, or it may be taken directly from the equipotential lines in Fig. 6, the point being $\frac{10}{50} = 20\%$ below E or at a height of 80% on the diagram. The value of this potential is 26%. The capacity of the antenna with respect to the ground plane may be calculated from equation 20.

$$C = \frac{.0388}{\log \frac{2D}{d}} = \frac{.0388}{\log \frac{2 \times 12 \times 100}{.5}} = .01055 \ \mu\text{f. per mile}$$

If we assume the antenna to be 1500 feet long its capacity will be $.01055 \times \frac{1500}{5280} = .003 \ \mu\text{f.}$

The capacity between the antenna wire and the power conductor is $C'' = C\frac{E'}{E} = .003 \times 26\% = .00078 \ \mu\text{f.}$ This is the useful or effective capacity of the coupling. The shunt capacity to ground, that is, the capacity representing no useful electric flux, is the total antenna capacity minus the useful coupling capacity or

$$C_g = C - C'' = .003 - .00078 = .00222 \ \mu\text{f.}$$

The power line of which P represents one conductor may be regarded as a line of infinite length. In this case the impedance which it presents at the sending end is its characteristic impedance. At frequencies of the order of 50,000 cycles which we are considering here the characteristic impedance is a practically pure resistance of approximately 400 ohms. We may therefore cut the exposed power conductor free from the rest of the line and replace the effect of the line by 400 ohms of non-inductive resistance as shown in Fig. 18.

The current flowing to ground through the 400 ohms terminating resistance is slightly less than the current flowing to ground when the power conductor is directly

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grounded. For simplicity we will assume the current passing through the 400 ohm resistance is determined by

$$I' = I\frac{E'}{E} = I \times .26 = .26 \text{ amps. per amp. in antenna.} \tag{34}$$

The error in the above assumption is not large enough to make any practical difference in our results. The current flowing through this 400 ohm resistance represents the useful current delivered to the power line. The energy delivered to the power line is

$$R' I'^2 = 400 \times .26^2 = 27.08 \text{ watts} \tag{35}$$

This is the useful energy delivered to the power line with one ampere in the antenna. This energy comes from the antenna and therefore the antenna input must contain an energy component equal to this, on the basis of the law of conservation of energy. The reaction of the power line on the antenna is then such as to introduce a resistance into the antenna circuit which when multiplied by the square of the antenna current, will equal the energy delivered to the power line. Thus

$$RI^2 = R'I'^2$$

$$\text{or } R = R'\left(\frac{I'}{I}\right)^2 \tag{36}$$

In this case

$$R = 400\left(\frac{.26}{1.}\right)^2 = 27.0 \text{ ohms}$$

This means that the power line under the conditions assumed, reacts into the antenna circuit like a resistance of 27 ohms and represents useful energy.

Antenna Transformation Radio Same as for a Step-Up Transformer

This transformation of the resistance on the line side into its equivalent on the antenna side of the system is exactly analogous to the usual method of converting a resistance on one side of a transformer to its equivalent on the other side by multiplying by the square of the ratio of transformation. If, instead of an antenna system, we had a transformer with a step-up ratio arranged to deliver .26 amps. to a 400 ohm resistance on its secondary with one ampere in its primary, the equivalent primary resistance would be

$$R = 400 \times (ratio)^2 = 400 \times \left(\frac{.26}{1.0}\right)^2 = 27 \text{ ohms}$$

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Hence the antenna acts exactly like a transformer having a step-up ratio of $\frac{1.0}{.26}=3.85$.

It is interesting to note that the ratio of transformation of this electrostatic transformer is determined by the potential ratio $\frac{E'}{E}$ of the power conductor with respect to the antenna. If we regard the system as a step-up transformer with the power line as a load on the secondary, the step-up ratio is the reciprocal of the potential ratio $\frac{E'}{E}$, that is, it is $\frac{E}{E'}$.

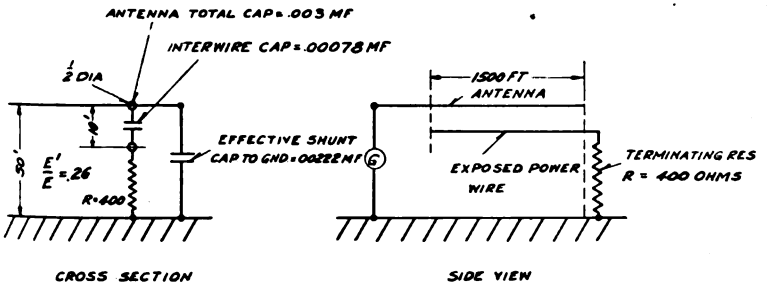

Fig. 18.—Antenna efficiency $=\frac{\text{Total Input Res.}}{\text{Useful Res.}}=\frac{27.0}{27.76}=97.5\%$

Since a power line represents a relatively high resistance as a load, a step-up ratio of voltage transformation is exactly what is wanted to couple a low resistance generating system to such a line.

In determining the efficiency of a transformer it is usual to arrive at the input by adding all the known losses to the output and dividing the output by the input. This practice may be followed in the case of antenna coupling system. The losses in the antenna system are due to dielectric absorption, radiation and the resistances of the primary and secondary conductors. The resistances of the primary and secondary are the resistances of the antenna wire and the resistance of the power conductor immediately under the antenna, since this part of the power line may be charged against the antenna as part of the coupling system.

Resistance of Antenna Conductors

The resistance of a copper conductor .5 inches in diameter at 50,000 cycles, as determined by the method recom-

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mended by the Bureau of Standards is 10.86 multiplied by its d-c. resistance which gives 2.47 ohms per mile of one wire. Tests on actual power lines show that the resistance per mile is much higher than the above figure. This is probably due to the effective resistance as measured including the various other losses, such as radiation and absorption, etc. The value of resistance as determined by test on a typical power line is approximately 8.0 ohms per mile of one wire and this value will be used in calculating the losses in the antenna. It is assumed that the antenna current is uniformly diminished toward the open end at which point it is of course zero. The effective resistance of a wire having such a current distribution is 1/3 the total resistance. If the current were distributed according to a sine law the effective resistance would be ½ the total resistance. The distribution is sufficiently close to the straight line law to be assumed correct in this case. The resistance of 1500 feet of antenna wire is 2.28 ohms. Assuming a straight line current distribution the effective antenna resistance is .76 ohms. The total antenna resistance is $.76 + 27.0 = 27.76$ ohms. The efficiency of the system is

$$\frac{useful\ res.}{total\ \ res.} = \frac{27.0}{27.76} = 97.5\%$$

Ground Losses Neglected

It will be observed in the foregoing that the ground is assumed to be a perfect conducting plane. In practice there are appreciable losses in the ground under the antenna and there are much more serious losses at the ground stake or system of ground stakes where the circuit makes contact with the ground. If we were interested in the practical case of efficiency of an antenna to ground these losses would have to be taken into consideration, and would make a reduction in the efficiency calculated. The ground stake losses, however, can be reduced almost as much as we like by subdividing the ground currents among a number of stakes. Such losses therefore are not chargeable directly to the mechanism of coupling, but rather to the existing conditions in any specific case. The case with which we have to deal in practice is usually one in which the ground does not enter, so the foregoing analysis will apply directly except that the dimensions of the coupling system are somewhat different.

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Ground losses are eliminated in much the same way as is done in radio by using a counterpoise in connection with an antenna. The antenna currents then flow entirely in metallic conductors and the electric field is not intercepted by the ground plane. The dielectric is air which is the best dielectric known so that the losses in the electric field due to absorption are very low as is assumed in the foregoing.

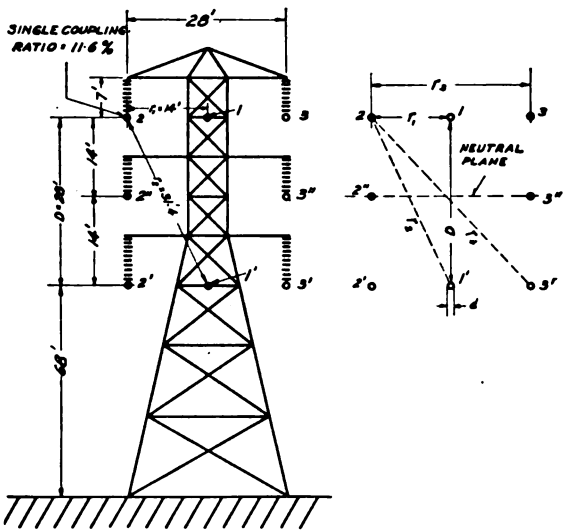

Fig. 19a. Fig. 19b.

Fig. 19.—Typical tower carrying 2-3 phase 132,000 volt circuits. Total coupling ratio $\frac{I'}{I} = 23.2\%$. Antenna efficiency 94%.

Phase to Phase Coupling

This method of coupling requires two coupling wires or antenna. These coupling wires are associated with the power conductors as intimately as possible. Fig. 19 shows a typical steel tower carrying two three phase power circuits. The position of the antenna is indicated at *1* and *1'*. These are arranged to be on a level with the top and bottom phase wires if possible. The antenna wires *1* and *1'* are connected to the transmitter loading coil so as to be at equal and opposite potentials with respect to ground at all times. By this arrangement charges of opposite polarity are induced in the top and bottom pairs of power conductors, which thus causes opposing currents to flow in these wires.

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Such a coupling system avoids the use of ground as a part of the circuit both in the antenna and in the line. The system is sometimes referred to as phase to phase, inter-phase or metallic circuit coupling.

As an example of phase-to-phase coupling, assume the distance between the antenna wires to be 28 feet with 14 feet between the antenna at the center of the arm, and the power conductor on the same level. If the conductors are all 300,000 cir. mils the coupling potential will be

$$\frac{E'}{E} = \frac{\log \frac{r_2}{r_1}}{\log \frac{2D}{d}} = \frac{\log \frac{31.4}{14.0}}{\log \frac{2 \times 12 \times 28}{.63}} = 11.6\%$$

The characteristic impedance of a power line having this size conductor and spacing is about 850 ohms per circuit. If we assume the two circuits to act independently, each would be terminated in its characteristic resistance of 850 ohms or they may be in parallel giving a resultant of 425 ohms. Since the coupling percentage of 11.6% applies to both circuits the total current appearing in the power lines will be 23.2% of the antenna current. The terminating resistance of 425 ohms reacts into the antenna side of the coupling system as

$$R = R' \left(\frac{I'}{I}\right)^2 = 425 \times (.232)^2 = 22.8 \text{ ohms}$$

Assuming as before 8 ohms per mile of single conductor as an approximation of the conductor resistance, the resistance of 1500 feet of antenna consisting of two wires is 4.56 ohms. Since the current is not constant throughout this length but diminishes from a maximum to zero at the open end, its effective resistance will be 1/3 of this or 1.52 ohms. The overall antenna efficiency will be

$$\frac{\text{useful Res. Com.}}{\text{total Res.}} = \frac{22.8}{22.8 + 1.52} = 94\%$$

Analysis Taking Reaction Between Wires Into Consideration

In the analysis up to this point no consideration has been given to the reaction of the charge on the power conductors on the charge on the antenna. These effects may be readily calculated by the method given by Maxwell. This method is based on the principle that the potential at any point is

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equal to the sum of the potentials produced independently by the several charges. As pointed out in connection with equation 15, the potential produced at a given point P by a charge Q above a conducting plane is equal to

$$E_p = 2\,q\,log\,\frac{r_2}{r_1}$$

This relation consists of a factor $2\,log\,\frac{r_2}{r_1}$ which multiplied by the charge q gives the potential at the desired point and may be written

$$E = qp \qquad (37)$$

The factor p is called a potential coefficient. If we have several charges such as q_1, q_2, q_3, all acting to produce a resultant potential at P and if each of the charges acting independently has a potential coefficient p_1, p_2, p_3 respectively the resultant potential will be

$$E_p = q_1\,p_1 + q_2\,p_2 + q_3\,p_3 \qquad (38)$$

Applying this principle to the case of a single antenna wire, Fig. 17 acting on a grounded power conductor, the resultant potential of the antenna wire is made up of two parts, first that produced by the charge on the antenna wire itself, and second by the effect of the charge on the grounded power wire. Using the usual notation

$$E_1 = q_1\,p_{1,1} + q_2\,p_{2,1} \qquad (39)$$

$$E_2 = q_2\,p_{2,2} + q_1\,p_{1,2} \qquad (40)$$

The first part of each subscript of the potential coefficient indicates the body on which the charge is located and the second part indicates the point or body on which the potential is induced.

In equation 39 the first term $q_1\,p_{1,1}$ gives the potential of the antenna wire due to its own charge q_1. The potential coefficient in this case is simply reciprocal of the antenna capacity i. e., $p_1 = \frac{1}{C} = 2\,log\frac{2D}{d}$ as in equation 19. The second term $q_2\,p_{2,1}$ gives the potential induced on the antenna by the charge q_2 on the power conductor. The potential coefficient in the case is $p_{2,1} = 2\,log\frac{r_2}{r_1}$ where r_1 and r_2 are the distances from the power wire to the antenna and its image respectively.

If we assume that the charge q_1 is positive and that the power conductor is grounded, its potential will be zero and

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the charge q_2 induced upon it will be negative. The equations may then be rewritten

$$E_1 = q_1\ p_{11} - q_2\ p_{21} \tag{41}$$

$$0 = -q_2\ p_{22} + q_1\ p_{12} \tag{42}$$

The values of the potential coefficients are:

$$p_{11} = p_{22} = 2\ log\ \frac{2D}{d} \quad p_{12} = p_{21} = 2\ log\ \frac{r_2}{r_1}$$

Substituting these values in equations 41 and 42

$$E_1 = q_1 \left(2\ log \frac{2D}{d}\right) - q_2 \left(2\ log \frac{r_2}{r_1}\right) \tag{43}$$

$$0 = -q_2 \left(2\ log \frac{2D}{d}\right) + q_1 \left(2\ log \frac{r_2}{r_1}\right) \tag{44}$$

From equation 44, we get the ratio of the charges on the two wires.

$$\frac{q_2}{q_1} = \frac{log\ \frac{r_2}{r_1}}{log\ \frac{2D}{d}} \tag{45}$$

From equation 18

$$\frac{E'}{E} = \frac{log\ \frac{r_2}{r_1}}{log\ \frac{2D}{d}}$$

Hence

$$\frac{q_2}{q_1} = \frac{E'}{E} \tag{46}$$

That is, the ratio of induced to inducing charge is exactly equal to the ratio of induced to inducing potential as assumed in equations 28, 29, and 30. Since the currents in the power line and antenna are proportional to the charges, it follows that $\frac{I'}{I}$ is just the same whether we take into account the reaction of the charge on the power conductor or neglect it. The ratio of transformation, which of course is the ratio of primary and secondary currents, is therefore unaltered by including the reaction of the charges on the power conductor. The equivalent primary resistance of the secondary load is therefore unaltered, which means that the efficiency of the antenna system works out to be correct even though we do neglect the reaction of the charge on the power conductor.

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Increase in Antenna Capacity

The actual antenna capacity, however, is slightly increased. Substituting the values of q_2 determined from the equation 45 in the equation 43, we get

$$E_1 = q_1 \left[2 \log \frac{2D}{d}\right] - \frac{q_1 \log \frac{r_2}{r_1}}{\log \frac{2D}{d}} \cdot 2 \log \frac{r_2}{r_1}$$

$$= q_1 2 \log \frac{2D}{d} - 2 q_1 \frac{\left[\log \frac{r_2}{r_1}\right]^2}{\log \frac{2D}{d}} \qquad (47)$$

The capacity of the antenna is measured by the charge required to raise its potential 1 volt, that is

$$C = \frac{q_1}{E_1} = \frac{1}{2 \log \frac{2D}{d} - 2 \frac{\left[\log \frac{r_2}{r_1}\right]^2}{\log \frac{2D}{d}}} \qquad (48)$$

It may be noted that the first term in the denominator is identical with the denominator of equation 19, which represents the antenna capacity without regard to the presence of the charge on the power conductor. The second term makes a correction for the effect of the charge on the power conductor. It is to be subtracted from the first term and therefore causes the total effective capacity to be somewhat greater than the capacity of the antenna wire alone.

The increase in capacity can be obtained by taking the ratio of the capacity including reactions as given in equation 48 to the capacity of the antenna wire alone given in equation 19. This gives

$$\frac{C'}{C} = \frac{2 \log \frac{2D}{d}}{2 \log \frac{2D}{a} - 2 \frac{\left[\log \frac{r_2}{r_1}\right]^2}{\log \frac{2D}{d}}} \qquad (49)$$

For the case given in the example shown in Fig. 18 for a ½-in. antenna wire, 50 feet above the ground plane and

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having a power conductor 10 ft. below it, the values are

$$\log \frac{2D}{d} = \log \frac{2 \times 100 \times 12}{.5} = 3.681$$

$$\log \frac{r_2}{r_1} = \log \frac{90}{10} = .955$$

Substituting these values in equation 49

$$\frac{C'}{C} = \frac{2 \times 3.681}{2 \times 3.681 - 2 \frac{.955^2}{3.681}} = 1.075 \qquad (50)$$

Hence the antenna capacity in this case is increased 7.5% by the presence of the power conductor 10 ft. below it.

Application to Double Circuit Line

The method of potential coefficients may be applied to the case of a double circuit power line having phase to phase antenna coupling. In this case we have a total of six power conductors to be considered. Under the conditions arising in practice where the antenna coupling wires are on the same level as the upper and lower pairs of power conductors, or are at least symmetrically arranged with respect to the middle power conductors, the case is greatly simplified by omitting the middle conductors entirely from consideration. This may be done since by symmetry the middle conductors are always located on the neutral plane, and therefore have no charge and hence do not enter the problem. The case may be further simplified by considering only the upper two power conductors and the upper antenna wire, the neutral plane being assumed to be a perfectly conducting ground plane. The potentials are then calculated by regarding the lower power conductors and lower antenna wire as images. These images are *real* in this case. Since the antenna wire is located symmetrically with respect to the upper two power conductors the currents and charges on these two wires must be equal. Referring to Fig. 19-b the potential equations are

$$E_1 = q_1 p_{11} - q_2 p_{21} - q_3 p_{31} \qquad (51)$$

$$0 = -q_2 p_{22} + q_1 p_{12} - q_3 p_{32} \qquad (52)$$

Since by symmetry

$$q_2 = q_3 \quad p_{11} = p_{22} \quad p_{12} = p_{21} = p_{31}$$

and $p_{11} = 2 \log \frac{2D}{d}$ $p_{12} = 2 \log \frac{r_2}{r_1}$ $p_{32} = 2 \log \frac{r_4}{r_3}$

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$$E_1 = q_1\left(2\log\frac{2D}{d}\right) - 2q_2\left(2\log\frac{r_2}{r_1}\right) \quad (53)$$

$$0 = -q_2\left(2\log\frac{2D}{d} + 2\log\frac{r_4}{r_3}\right) + q_1\left(2\log\frac{r_2}{r_1}\right) \quad (54)$$

The last equation gives the ratio of transformation, that is, the ratio of charge induced on the grounded power conductor to the charge on the antenna.

$$\frac{q_2}{q_1} = \frac{\log\frac{r_2}{r_1}}{\log\frac{2D}{d} + \log\frac{r_4}{r_3}} \quad (55)$$

Since there are two power conductors, each of which has a charge q_2, the over-all ratio of transformation is just twice the above or

$$\frac{q'_2}{q_1} = \frac{2q_2}{q_1} = \frac{2\log\frac{r_2}{r_1}}{\log\frac{2D}{d} + \log\frac{r_4}{r_3}} \quad (55a)$$

It will be noted by comparison with equation 18 that the ratio is identical with the potential ratio $\frac{E'}{E}$ for the conductors involved *except* that the denominator is increased by the term $\log\frac{r_4}{r_3}$. The charge induced on conductor No. 2 therefore is slightly less by reason of the presence of power conductor No. 3. The greater the distance between the power conductors the less this effect and the more nearly they act like independent circuits.

Substituting the value of q_2 from equation 55 in 53

$$E_1 = 2q_1\log\frac{2D}{d} - 4q_1\frac{\left(\log\frac{r_2}{r_1}\right)^2}{\log\frac{2D}{d} + \log\frac{r_4}{r_3}} \quad (56)$$

The capacity of the antenna wire to the neutral plane taking into consideration the reaction of the charges on the two power conductors above the neutral plane, is

$$C = \frac{q_1}{E_1} = \frac{1}{2\log\frac{2D}{d} - 4\frac{\left(\log\frac{r_2}{r_1}\right)^2}{\log\frac{2D}{d} + \log\frac{r_4}{r_3}}} \quad (57)$$

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If we omit the entire second term in the denominator the equation gives the capacity of the antenna wire to the neutral plane, neglecting the reaction of the power conductors, as in equation 19. The effect of this reaction is to somewhat increase the antenna capacity.

The interwire capacity, that is, the capacity between the antenna wire and the two associated top power conductors is

$$C'' = C\,\frac{2q_2}{q_1} = C\,\frac{2\log\frac{r_2}{r_1}}{\log\frac{2D}{d} + \log\frac{r_4}{r_3}} \qquad (58)$$

where C is the total or effective antenna capacity as determined by equation 57. This reduces to

$$C'' = \frac{2\log\frac{r_2}{r_1}}{2\left(\log\frac{2D}{d}\right)^2 + 2\log\frac{2D}{d} \times \log\frac{r_4}{r_3} - 4\left(\log\frac{r_2}{r_1}\right)^2}$$

The last term in the denominator corrects for the effect of the reaction of the power conductors. If this term is omitted the equation reduces to the interwire capacity obtained by neglecting the reaction of the charges on the power conductor. These equations may be applied to the case of phase-to-phase antenna coupling previously worked out. They now take into consideration the reaction of the charges on the power conductors. Referring to Figures 19-a and 19-b

$$\log\frac{2D}{d} = \frac{2 \times 12 \times 28}{.63} = 3.0273$$

$$\log\frac{r_2}{r_1} = \log\frac{31.4}{14} = .350 \text{ and } \log\frac{r_4}{r_3} = \log\frac{39.6}{28.} = .1507$$

The ratio of transformation is

$$\frac{q'_2}{q_1} = \frac{2\log\frac{r_2}{r_1}}{\log\frac{2D}{d} + \log\frac{r_4}{r_3}} = \frac{2 \times .350}{3.027 + .150} = 22.0\%$$

It will be noted that this is changed but very slightly from the previous value of 23.1% obtained by neglecting the second term in the denominator. Hence in this case the power conductors are so far apart that they have but little effect on each other.

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Using the new value of 22.0% as the correct ratio of transformation, we get the equivalent primary resistance due to the secondary resistance of 425 ohms, of

$$R = R'\left(\frac{E'}{E}\right)^2 = 425 \times .22^2 = 21 \text{ ohms}$$

The resistance of the antenna wire is 1.52 ohms, therefore the efficiency of the circuit is

$$\frac{21.0}{21.0 + 1.52} = 93.5\%$$

the previous efficiency was 94%

The capacity of an antenna wire to the neutral plane is

$$C = \frac{1}{2 \log \frac{2D}{d}}$$

The ratio of the total effective capacity of an antenna wire to the simple capacity of the single wire to the neutral plane is obtained by combining this with the equation 57, which gives

$$\frac{C'}{C} = \frac{2 \log \frac{2D}{d}}{2 \log \frac{2D}{d} - \frac{4\left(\log \frac{r_2}{r_1}\right)^2}{\log \frac{2D}{d} + \log \frac{r_4}{r_3}}} \qquad (59)$$

Applying this to the line under consideration we get

$$\frac{C'}{C} = \frac{2 \times 3.027}{2 \times 3.027 - \frac{4 \times .35^2}{3.027 + .150}} = 1.025$$

Where $\log \frac{2D}{d} = 3.027$, $\log \frac{r_2}{r_1} = .35$ and $\log \frac{r_4}{r_3} = .150$

Hence the antenna capacity is increased by only 2.5%.

The interwire capacity, that is, the effective capacity between the antenna and the power line as determined by equation 58 is

$$C'' = C \frac{2q_2}{q_1} = C \times 1.025 \times 22\%$$

The value of C, the capacity of the free antenna wire to the neutral plane is

$$C = \frac{.0388}{\log \frac{2D}{d}} = \frac{.0388}{3.027} = .0128 \ \mu\text{f. per mile of one wire.}$$

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Assuming the antenna length to be 1500 feet this gives

$$C = .0128 \times \frac{1500}{5280} = .00364 \ \mu f.$$

$$C' = .00364 \times 1.025 = .00374 \ \mu f.$$

Substituting this value for C in equation 58 we get

$$C'' = .00364 \times 1.025 \times 22\% = .00082 \ \mu f.$$

This is the total capacity which couples the antenna wire to the upper two power conductors. The capacity to each conductor is one half of this or $.00041$ μf.

It is important to note that the presence of the group of wires represented by a double circuit power line increases the antenna capacity by so small a value as 2.5%. It is also of interest to note that this increase in capacity is less than in the previous case of phase-to-ground coupling where the increase is 7½%. This would seem rather inconsistent, but a study of the equations involved will show that the effect is due to the low coupling ratio in the phase-to-phase case, which is only 11% to each wire, whereas in the phase-to-ground case the coupling ratio is 26%. This striking difference in coupling ratio is the result of the very great change in distances involved. The value of D in the phase-to-ground case is the distance between the conductor and its image, or 100 feet, while in the phase-to-phase case, it is only 28 feet.

Effect of Line Terminating Resistance

The analysis up to this point, while taking into consideration all the capacity reactions between the antenna and the various wires of the power line, it has been assumed that the induced current in the power line due to the current in the antenna is the same as the current in the power line, when the power line is shorted to ground in the case of phase-to-ground coupling. In other words, it is assumed that the effect of the line characteristic impedance is negligible in comparison with the impedance of the interwire capacity. That this is true may be demonstrated by calculating the current through the terminating impedance as indicated in Figs. 20-A and 20-B, which shows the relations of the various elements in an antenna system. It is assumed that the line characteristic resistance is 850 ohms. This is represented in the section 2, 3 of the diagram. The total an-

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tenna capacity involving all reactions for the double circuit line previously calculated is .00374 μf. Of this, .00082 μf. is interwire capacity. The capacity of each power conductor to the neutral plane is .00374, but when two wires are tied in parallel at the given spacings the capacity of each wire is only 95% of the original, or both wires have together 1.90% of one wire. Thus the capacity of the two power conductors to the neutral plane is $1.90 \times .00374 = .00712$, of which .00082 is the interwire capacity between them and the antenna. The net capacity of the two power conductors to the

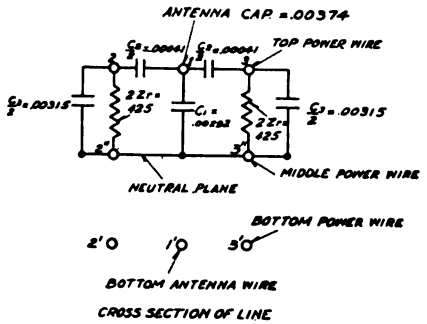

Fig. 20a.

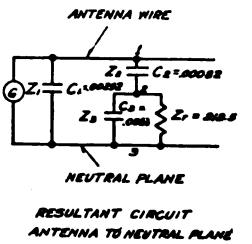

Fig. 20b.

Fig. 20.—Complete analysis of phase-to-phase antenna coupling to double circuit power line efficiency = 92.2%.

neutral plane which shunts the line terminating resistance is $.00712 - .00082 = .0063$ μf. This capacity is in shunt with the characteristic impedance of the two lines in parallel to neutral, or 212.5 ohms. The interwire capacity is shown by the condenser in the link 1, 2. The capacity of the antenna-to-ground shunting the whole system is $.00374 - .00082 = .00292$ μf.

The impedance of the parallel mesh 2, 3 is

$$Z_{2,3} = \frac{Z_3 Z_r}{Z_3 + Z_r} \qquad (60)$$

The impedance of the total path from 1 to 3 which includes the interwire capacity is

$$Z_{13} = Z_{12} + Z_{23} = Z_2 + \frac{Z_3 Z_r}{Z_3 + Z_r} \qquad (61)$$

The current passing from 1 to 3 by way of the interwire capacity is

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$$I_{13} = \frac{E}{Z_{13}} = \frac{E}{Z_2 + \frac{Z_3 Z_r}{Z_3 + Z_r}} \tag{62}$$

The drop across the resistance is

$$E_r = I_{13} Z_{23} = \frac{E}{Z_2 + \frac{Z_3 Z_r}{Z_3 + Z_r}} \times \frac{Z_3 Z_r}{Z_3 + Z_r} \tag{63}$$

$$I_r = \frac{E_r}{Z_r} = \frac{E}{Z_r} \frac{\frac{Z_3 Z_r}{Z_3 + Z_r}}{Z_2 + \frac{Z_3 Z_r}{Z_3 + Z_r}} \tag{64}$$

The total current in the antenna is

$$I = \frac{E}{Z_1} + I_{13} = \frac{E}{Z_1} + \frac{E}{Z_2 + \frac{Z_3 Z_r}{Z_3 + Z_r}} \tag{65}$$

$$\frac{I_r}{I} = \frac{1}{Z_r} \frac{\frac{Z_3 Z_r}{Z_3 + Z_r}}{\frac{1}{Z_1} + \frac{1}{Z_2 + \frac{Z_3 Z_r}{Z_3 + Z_r}}} \tag{66}$$

If we neglect the term $\frac{Z_3 Z_r}{Z_3 + Z}$ which is small in comparison with Z_2 we get

$$\frac{I_r}{I} = \frac{Z'}{Z_2} \times \frac{Z_3}{Z_3 + Z_r}$$

where Z' is the total antenna impedance.

This relation will enable us to calculate the current going into the power line per ampere of antenna current. $\frac{I_r}{I}$ is the coupling ratio and can be used in determining the final overall efficiency.

Applying this to the example of the double circuit line we get

$$Z' = \frac{1}{j\omega C} = -j \frac{1}{6.28 \times 50{,}000 \times .00374 \times 10^{-6}}$$

$$= -j\,853 \text{ ohms}$$

$$Z_2 = \frac{1}{j\omega C_2} = -j \frac{1}{6.28 \times 50{,}000 \times .00082 \times 10^{-6}}$$

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$$= -j\,3900 \text{ ohms}$$

$$Z_3 = \frac{1}{j\omega C_3} = -j\,\frac{1}{6.28 \times 50{,}000 \times .0063 \times 10^{-6}}$$

$$= -j\,507 \text{ ohms}$$

$$Z_r = \frac{425}{2} = 212.5$$

$$\frac{I_r}{I} = \frac{Z'}{Z_2} \cdot \frac{Z_3}{Z_3 + Z_r} = \frac{-j853\,(-j507)}{(-j\,3900)\,(212.5 - j\,507)}$$

$$\frac{I_r}{I} = \frac{853 \times 507}{3900 \times 546} = 20.3\%$$ which is the transformation ratio with all reactions considered. The terminating resistances on the two power lines in parallel transformed into the equivalent primary or antenna resistance is

$$R = 425 \times \left(\frac{I_r}{I}\right)^2 = 425 \times (.203)^2 = 17.5 \text{ ohms.}$$

The total antenna resistance $= 17.5 + 1.52 = 19.02$ ohms.

The resultant efficiency is $\frac{17.5}{19.02} = 92.2\%$ taking everything into account.

Conclusion Regarding Antenna Efficiency

From the foregoing analysis it is quite apparent that the present popular idea that antenna coupling is inefficient, is not well founded. The analysis shows that with the coupling distances ordinarily obtainable in practice, the efficiency of antenna coupling is so high that the coupling losses do not seriously affect the energy actually delivered to the line by the transmitter.

Of the two methods of coupling available, namely, antenna coupling and condenser coupling, there is little preference either way on the score of coupling efficiency. The decision as to whether antenna coupling or condenser coupling should be used in any given case, must be made on altogether different grounds than that of coupling efficiency.

For high-voltage lines, antenna coupling is in general to be preferred. This preference is because of the fact that with antenna coupling the dielectric is air, which is the best insulation obtainable. There is no progressive deterioration to be feared in a condenser having an air dielectric. The coupling wires are out in the open and always accessible for inspection. The method of supporting the coupling

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wires is exactly the same as that used in regular line construction, and is therefore something with which a power man has much experience and in which he has great confidence. Antenna coupling is in general much safer than condenser coupling, and for high voltage lines it is less expensive.

There are circumstances under which antenna coupling is at a disadvantage. These considerations are mechanical rather than electrical. Sometimes the lines leaving a station swing immediately across a river and the supporting structures are such that no provision can be made for carrying the necessary antenna wires, without seriously altering or rebuilding the steel work. In such a contingency coupling condensers find a proper application. The physical arrangement of the high tension apparatus, and the manner in which the lines leave the station sometimes make it difficult properly to locate antenna wires. Sometimes an adjacent circuit, in which we have no interest, will be exposed to the inducing field of the antenna and thus rob us of useful communication energy. For low voltage lines, the cost of condenser coupling is often less than for antenna coupling. It might be said in general that where for mechanical reasons there are objections to mounting antenna coupling wires, coupling condensers have a legitimate application.

Propagation of Alternating Currents on an Infinite Line

In the study of lines, especially when transmitting high-frequency currents, their properties are more easily understood when the length is made infinite. The behavior of an infinite line is much more simple than that of a finite line, because it is not complicated by reflections from the end. With such a line all disturbances originating at the sending end are propagated down the line and can never return to combine with, and interfere with the advancing waves coming from the sending end generator.

When alternating currents are propagated down an infinite line, the amplitude of the voltage vector is progressively reduced as the distance from the sending end is increased. The percentage decrease in amplitude per mile is a constant throughout the length of the line. The amplitude

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of the voltage vector at a point x miles from the sending end may be expressed as

$$E_x = E_1 e^{-ax} \qquad (67)$$

The reduction in amplitude per mile is given by the factor e^{-a}. It is the attenuation ratio per mile of line. The exponent a is the attenuation constant.

In addition to being progressively reduced in amplitude, the phase of the voltage vector is progressively shifted backwards as the distance from the sending end increases. That is, the vector representing the potential between wires at a given point on the line, differs in phase from the vector representing the potential at a point farther down the line. The potential at the more distant point always lags behind the potential at a point nearer the sending end. This phase displacement is the same for each mile of line, that is, it is a constant. The phase lag per mile is called the wavelength constant. The phase lag of the voltage vector at a point x miles from the sending end is

$$\theta = B x \qquad (68)$$

where B is the wave length constant.

The attenuation of the voltage or current vectors as well as the retardation in phase, can be represented diagrammatically as in Fig. 21. The line conductors are represented by A, B and C, D. The potential between lines at the point A is represented by the length of the vector E_1, which is drawn vertically as a position of reference. At points representing one mile intervals circles are drawn whose radii are progressively reduced by the factor e^{-a}. Thus the radius of the circle at the end of the first mile is $E_1 e^{-a}$. The circle at the end of the second mile is $E_1 e^{2-a}$ etc. The circle at the distance x is defined by equation 67. The voltage vector at the end of the first mile is drawn so as to lag behind E_1 by the angle B. The voltage vector at the end of the second mile lags by the angle $2B$ etc.

The potential at the point x may be completely represented by combining equations 67 and 68 thus

$$E_x = E_1 e^{-ax} \cdot e^{-jBx} \qquad (69)$$

The factor e^{-jBx} is an operator $(\cos Bx - j \sin Bx)$ which rotates the vector it acts on through the angle $(-Bx)$. It performs the operation required by equation 68 in that it locates the vector E_x behind E_1 by the angle Bx. The whole expression therefore gives the vector E_x both in magnitude

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and in phase with respect to E_1. Adding the exponents in equation 69 we get

$$E_x = E_1 e^{-(a+jB)x} = E_1 e^{-Px} \tag{70}$$

where $P = a + jB$. The quantity P is known as the propagation constant. It may be demonstrated to be*

$$P = \sqrt{R + j\omega L}\ \sqrt{G + j\omega C} \tag{71}$$

where R, L, G, and C are the fundamental constants of the line per unit length.

The current in an infinite line is attenuated just like the

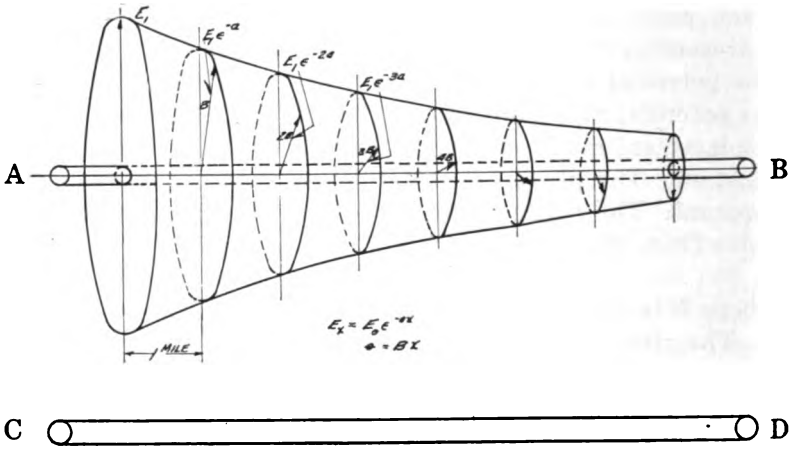

Fig. 21.—Diagram showing the phase and magnitude of potential at various points along a line.

voltage. The current at any point in the line without regard to phase is expressed by an equation similar to equation 67.

$$I_x = I_1 e^{-ax} \tag{72}$$

If we insert an ammeter at the sending end and another at a point in the line x miles away, we can use the ratio of these two readings to determine the attenuation constant a. Thus,

$$\frac{I_2}{I_1} = e^{-ax} \tag{73}$$

from which we get the attenuation constant,

$$a = \frac{log_{10} \frac{I_1}{I_2}}{.4343\ x} \tag{74}$$

This equation is of importance since it enables us to deter-

*See "Propagation of Electric Currents" by Fleming.

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mine the attenuation constant by making observations on the line.

Characteristic Impedance

The current entering an infinite line may be expressed in terms of the sending end applied voltage and the impedance which the line offers to the sending end generator. Thus,

$$I_1 = \frac{E_1}{Z_0} \tag{75}$$

where Z_0 is the vector impedance of the line viewed from the sending end. If a piece of the given infinite line were cut off and the sending end generator connected to the remainder, the current flowing into the line will still be defined exactly by equation 75. The impedance which the line presents to the sending end generator will still be exactly Z_0 as before since the line is infinite. It is a property of an infinite line, therefore, that it presents a constant impedance to the sending generator irrespective of the point where the generator may be connected. This constant impedance is called the "characteristic impedance" of the line.

If a piece of an infinite line is cut off and the open end closed by an impedance equal to the characteristic impedance, it will act on the sending generator exactly like the corresponding infinite line. Under this condition the currents at the sending and receiving ends may be measured, and the attenuation constant determined by the relations expressed in equations 73 and 74. It is important to note that the attenuation constant can be determined directly from the ratio of sending and received currents *only in case the line is infinite,* or has been converted into its equivalent infinite line by being terminated in its characteristic impedance. It is therefore necessary to determine accurately the characteristic impedance of the line in order that it shall be properly terminated before making measurements of current ratio.

The usual method of determining the characteristic impedance of an actual line under test is to measure the impedance of the line with the distant end open, and again with the distant end short. The characteristic impedance is then the square root of the product of these open and short impedances thus,

$$Z_0 = \sqrt{Z_{sh} \times Z_{op}} \tag{76}$$

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*The analytical basis of this equation is clearly set forth in standard works on the subject. This method is theoretically applicable to lines at all frequencies. It is quite satisfactory at low frequencies, but is not quite so well suited to work at high frequencies. Unless great precautions are taken serious errors are likely to occur which carry with it corresponding errors in the attenuation ratio.

Simple Method for Determining Characteristic Impedance

A much more simple and more reliable method has been developed for use on lines at high frequency. As pointed out above the impedance looking into a finite line from the sending end, will be the characteristic impedance, provided the distant end is closed through an impedance equal to the characteristic impedance. If therefore an impedance can be found such that when bridged across the open end of a finite line, the measured sending end impedance is equal to it, then this impedance is the characteristic impedance of the line. The method then is to place a trial impedance across the open end of the line under test, and measure the impedance at the sending end. The terminating impedance is varied until the impedance measured at the sending end is equal to it. The terminating impedance is then the characteristic impedance of the line and the ratio of received to sending end current gives the attenuation ratio correctly.

For work on overhead lines at high frequencies, this method is easily carried out in practice, since in this case the characteristic impedance is always a simple non-inductive resistance. This can been seen by inspection of the fundamental equation for characteristic impedance in terms of the fundamental line constants,

$$Z_0 = \frac{\sqrt{R + j\omega L}}{\sqrt{G + j\omega C}} \qquad (77)$$

At high frequencies the value of R is always negligible when compared to ωL and G is negligible when compared to ωC. The equation therefore reduces to

$$Z_0 = \sqrt{\frac{L}{C}} \qquad (78)$$

Since the j component does not appear, the quantity is a pure resistance. Hence for work at high frequencies on open

*See "Propagation of Alternating Currents," by Fleming.

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wire lines, it is known beforehand that the characteristic impedance is a resistance. The trial terminating impedance, therefore, is a simple non-inductive resistance. The work of measurement at the sending end is also simplified since it is known that when the correct adjustment is reached, the sending end impedance will contain no reactive com-

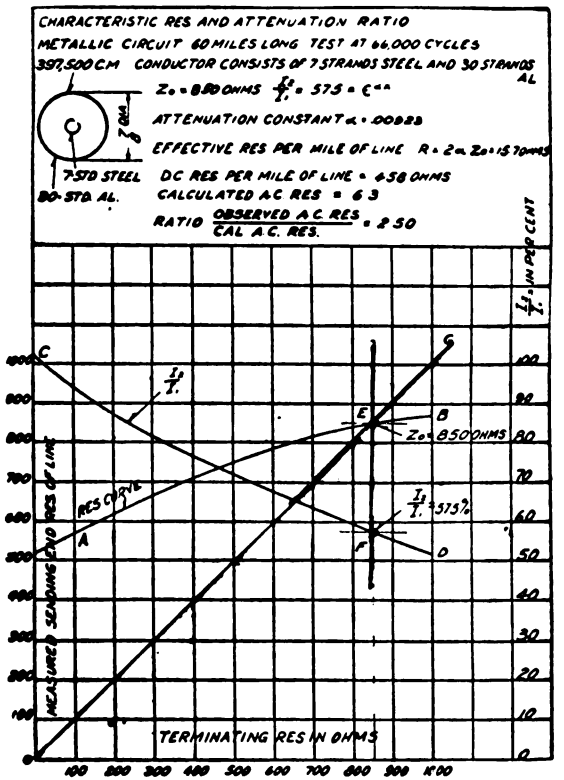
CHARACTERISTIC RES AND ATTENUATION RATIO
METALLIC CIRCUIT 60 MILES LONG TEST AT 66,000 CYCLES
397,500 CM CONDUCTOR CONSISTS OF 7 STRANDS STEEL AND 30 STRANDS AL
Z₀ = 850 OHMS
ATTENUATION CONSTANT α = .00088
EFFECTIVE RES PER MILE OF LINE R = 2α Z₀ = 1.5 OHMS
DC RES PER MILE OF LINE = 4.58 OHMS
CALCULATED A.C. RES = 6.3
RATIO OBSERVED A.C. RES / CAL A.C. RES. = 2.50
7 STD STEEL
30 STD AL.
I₀/I₀ IN PER CENT
MEASURED SENDING END RES OF LINE
RES CYCLE
Z₀ = 850 OHMS
I₀/I₀ = 57.5%
TERMINATING RES IN OHMS
100 200 300 400 500 600 700 800 900 1000

Fig. 22.

ponent. Hence only the resistance component of the sending end impedance need be measured.

In actually determining the characteristic impedance of a line under test a series of values of terminating resistance are used. For each value of terminating resistance the corresponding sending end resistance is measured. These data are then plotted in the form of a curve as shown in Fig. 22. The terminating resistances are plotted as abscissae and the measured sending end resistances are plotted as ordinates.

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The smooth curve A, B is then drawn through these points. The characteristic resistance is represented by a point on this curve whose coordinates are equal. This point may be determined by drawing the line $O\ G$ so as to bisect the right angle at O. The coordinates of any point in this bisector are equal. Hence the intersection of $O\ G$ with the resistance curve A, B determines the point "E", whose abscissas or ordinates are each equal to the characteristic impedance. The characteristic impedance for the line shown in Fig. 22 is 850 ohms.

In taking the data for the resistance curve, ammeters are placed at the sending and receiving ends. Readings of these meters are taken for each value of terminating resistance. The ratio of these currents $\frac{I_2}{I_1}$ is then plotted as curve C, D Fig. 22. The received currents for any value of terminating resistance can then be picked off this curve. The ratio of received to sending current for the case where the line is terminated in its characteristic impedance, is determined by drawing a vertical line through E. Its intersection F with the curve C, D gives the attenuation ratio. The value of the attenuation ratio for the line under consideration is 57.5%.

The line to which the data in Fig. 22 applies is a single circuit three phase line, whose configuration is given in Fig. 23. The test was made on a metallic circuit formed by using the two outside wires 1 and 3. The middle wire was open at both ends during the test. The frequency used was 66 kc. The power conductors were 397,500 cir. mil. aluminum steel reinforced. They consist of thirty strands of aluminum and seven strands of steel, the outside diameter being $\frac{7}{8}$ inch.

Apparatus Used in Line Measurements

The method of making the resistance measurements from which the curves in Fig. 22 were plotted is very simple. The only standardized apparatus required is a resistance box and an ammeter at each end of the line. The method is that commonly used in making resistance measurements at ratio frequencies, and is generally known as the resistance variation method. The arrangement is shown diagrammatically in Fig. 24. The measuring system consists

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of a condenser C and inductance L, a standard resistance box R, and a calibrated ammeter A connected in series to the hinges of a double throw switch. When this switch is thrown one way the line is introduced in series with the measuring circuit. When the switch is thrown the other way the line is cut out and the measuring circuit closed on itself. Energy for making the measurements is obtained by inductively coupling an oscillator to the inductance as indicated at M. The resistance of the measuring circuit is determined by throwing the double throw switch on short and adjusting the measuring circuit to resonance by the variable condenser. The coupling is adjusted so that a good reading preferably full scale is obtained on the ammeter. Enough resistance is then cut in to reduce the current to one half. The inserted resistance is then equal to the resistance already in the circuit. This gives the resistance of the measuring circuit. The double throw switch is then thrown on the line and the condenser adjusted for resonance. The resistance is again measured as before. This resistance is the total of the line plus the measuring circuit. By subtracting the resistance of the measuring circuit we get the resistance of the line. These measurements may be made very quickly and with considerable accuracy. The results obtained are much more reliable than those taken by more expensive equipment using the open and short impedance method. In the determination of characteristic resistance and attenuation by this method the frequency does not enter the observations or calculations. This eliminates one source of possible error.

The impedance of the line may be determined by the same apparatus if the condenser C is calibrated, and the frequency of the test is known. When the switch is thrown on short and the measuring circuit tuned to resonance, the reactance of the condenser C is just equal to the reactance of the inductance L. When the switch is thrown on the line and the condenser readjusted to resonance, the reactance of the condenser plus the reactance of the line must still be just equal to the reactance of the inductance. Hence if X_1 is the reactance of the condenser when tuning the measuring circuit alone, and X_2 is the reactance of the condenser when the measuring circuit contains the line in series we have,

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$$X_1 = X_2 + X'$$
$$\text{or } X' = X_1 - X_2 \qquad (79)$$
$$= \frac{1}{\omega C_1} - \frac{1}{\omega C_2}$$

where X' is the reactance of the line and C_1 and C_2 are the values of condenser capacity under the first and second conditions respectively. By this method the reactance component of the line can be determined for each frequency and for each condition of termination.

Attenuation Constant at High Frequencies

The attenuation constant is the real part of the propagation constant P as defined by equation 71. At high frequencies such that R is negligible in comparison with ωL the first radical may be expanded by the binomial theorem. The real part of the resulting expansion may be expressed as

$$a = \frac{R}{2}\sqrt{\frac{C}{L}} \qquad (80)$$

By comparing with equation 78 we can substitute for $\sqrt{\frac{C}{L}}$ its value $\frac{1}{Z_0}$. The relation then becomes

$$a = \frac{R}{2}\,\frac{1}{Z_0} \qquad (81)$$
$$\text{or } R = 2\,a\,Z_0$$

This equation is quite useful in making experimental studies of open wire lines at high frequencies. By the foregoing methods the attenuation constant a and the characteristic impedance Z_0 are determined readily by direct observation. Equation 81 enables us to use this data to calculate the effective resistance per mile of line. This is very useful since it is not possible to make direct measurement of line resistance at high frequencies owing to the distributed inductance and capacity. This effective resistance includes losses of all kinds which may be present in the line, in addition to the actual resistance losses in the conductors themselves. The conductor resistances alone may be calculated with accuracy from the tables and methods given by the Bureau of Standards. A comparison of the effective high frequency resistance of the line as observed with the calculated high frequency resistance of the conductors, gives data of considerable value.

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TESTS ON 397,500-CIR. MIL ALUMINUM CONDUCTORS.

In the case of the line whose data are plotted in curve Fig. 22 the characteristic impedance was found to be 850 ohms, and the ratio of received to transmitted current was .575 thus

$$e^{ax} = \frac{1}{.575}$$

$$.4343\ ax = log_{10} \frac{1}{.575} = .24033$$

The line on which the measurements were made was 60 miles long hence $x = 60$ which gives

$$a = \frac{.24033}{.4343 \times 60} = .00923$$

This is the attenuation constant of the line at the frequency of 66,000 cycles. The effective resistance of the line per mile is

$$R = 2a\ Z_o = 2 \times .00923 \times 850 = 15.7\ ohms$$

The conductor of which this line was made consists of 30 strands of aluminum, giving a total aluminum cross section 397,500 cir. mils. These aluminum strands were assembled around a steel core of 7 strands, the total outside diameter being 7/8 inch. The d-c. resistance of the aluminum part of the cable is .458 ohms. The ratio of measured effective a-c. resistance to d-c. resistance is $\frac{15.7}{.458} = 34.3$.

The ratio of a-c. to d-c. resistance calculated for this conductor by the method of the Bureau of Standards is

$$\frac{R}{R_o} = 13.69$$

where R is the a-c. resistance and R_o is the d-c. resistance. Using .458 ohms as the d-c. resistance R_o in the above we get the a-c. resistance at 66,000 cycles to be

$$R = .458 \times 13.69 = 6.3\ ohms$$

The ratio of the observed a-c. resistance of the circuit per mile to the calculated a-c. resistance of the conductors alone is $\frac{15.7}{6.3} = 2.50$. Hence there are losses in the circuit such that the effective resistance is 2.50 times as great as the resistance known to exist in the conductors themselves. These extra losses then amount to 150% of the losses in the conductors. By reference to Fig. 23 it will be seen that

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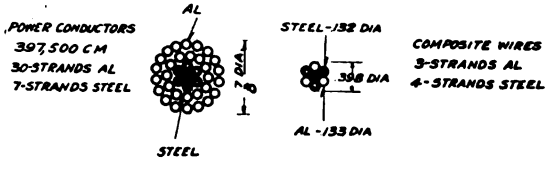
there are two insulated conductors twelve feet above the level of the power conductors whose resistances were being measured. These conductors are transposed with respect to the power line every five miles. Currents induced in these conductors from the currents flowing in the power line probably account for some of the losses in excess of the con-



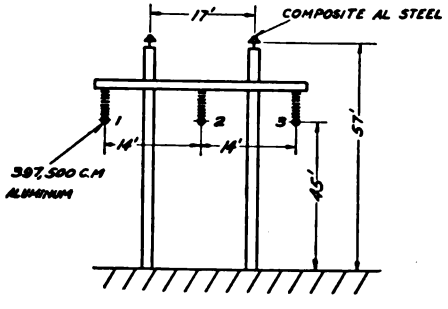


Fig. 23.

ductor resistance losses. In addition to this about 0.9% of the electric field between conductors 1 and 3 is intercepted by the ground plane. There are thus currents in the ground due to stray field from the line. It is of interest to note that the total value of the stray losses is not greater than 150% of the conductor losses. The value of Z_0 the line characteristic resistance may be determined by calculating the value of inductance and capacity per mile from the physical dimensions of the line thus,

$$L = 14.8 \times 10^{-4} \log_{10} \frac{2D}{d} \text{ henrys per loop mile.}$$

$$C = \frac{.01941}{\log_{10} \frac{2D}{d}} \ \mu f. \text{ per mile between wires}$$

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$$Z_0 = \sqrt{\frac{L}{C}} = \sqrt{\frac{14.8 \times 10^{-4} \log \frac{2D}{d}}{\frac{.01941 \times 10^{-6}}{\log \frac{2D}{d}}}}$$

$$= 276 \log_{10} \frac{2D}{d} \text{ ohms}$$

For the line shown in Fig. 21 using the outside power conductors 1 and 3 this becomes

$$Z_0 = 276 \times \log \frac{2 \times 12 \times 28}{.875} = 796 \text{ ohms}$$

The measured characteristic impedance was 850 ohms which is 6.5% higher than the calculated value.

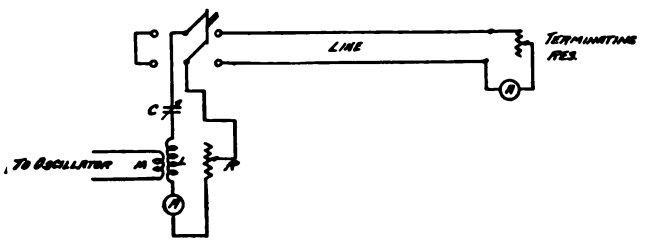

Fig. 24.—Apparatus and circuit used in making line measurements.

Velocity of Propagation

The velocity of propagation at high frequencies may be determined from

$$V = \frac{1}{\sqrt{L C}} \tag{82}$$

Comparing this with the relation

$$Z_0 = \sqrt{\frac{L}{C}}$$

it will be apparent that if the velocity were known and the characteristic impedance Z_0 were also known, the values of L and C would be determined thus

$$L = \frac{Z_0}{V} \tag{83}$$

and $$C = \frac{1}{Z_0 V} \tag{84}$$

Thus in the case of the line Fig. 22 in question, assuming

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the velocity to be 186,000 miles per second we get

$$L = \frac{Z_0}{V} = \frac{850}{186{,}000} = 4.57 \text{ milli-henrys per mile.}$$

$$C = \frac{1}{Z_0 V} = \frac{1}{850 \times 186{,}000} = .00633 \text{ MF. per mile}$$

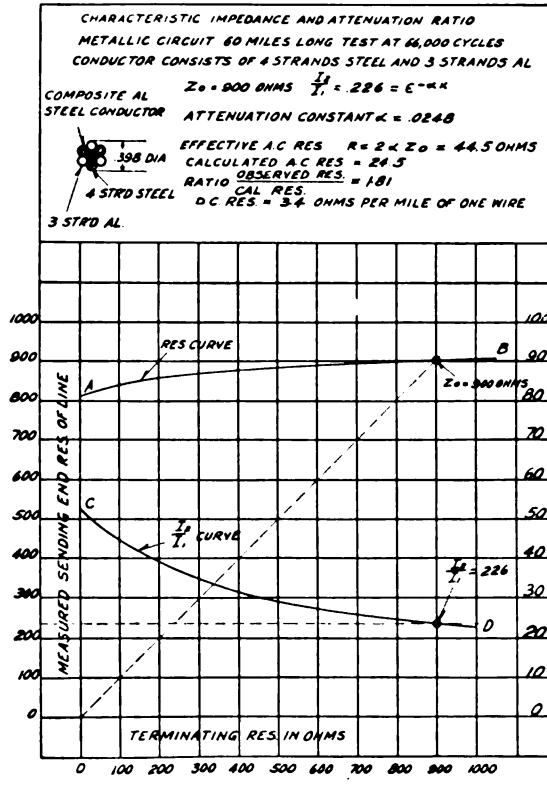

Fig. 25.

The value of L calculated from the dimensions of the line are

$$L = 14.8 \times 10^{-4} \log \frac{2D}{d}$$

$$= 14.8 \times 10^{-4} \times 2.8853 = 4.25 \text{ milli-henrys}$$

and the value of C is

$$C = \frac{.0194}{\log \frac{2D}{d}} = \frac{.0194}{2.8853} = .00672 \ \mu f.$$

Hence the ratio of the value of L deduced from observation

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of Z_0 to that calculated from the dimensions of the line is

$$\frac{L \text{ observed}}{L \text{ calculated}} = \frac{4.57 \times 10^{-3}}{4.25 \times 10^{-3}} = 107.5\%$$

and

$$\frac{C \text{ observed}}{C \text{ calculated}} = \frac{.00633}{.00672} = 94.2\%$$

Thus the observed inductance is 7.5% high and the observed capacity is 5.8% low. If we substitute in equation 82 the

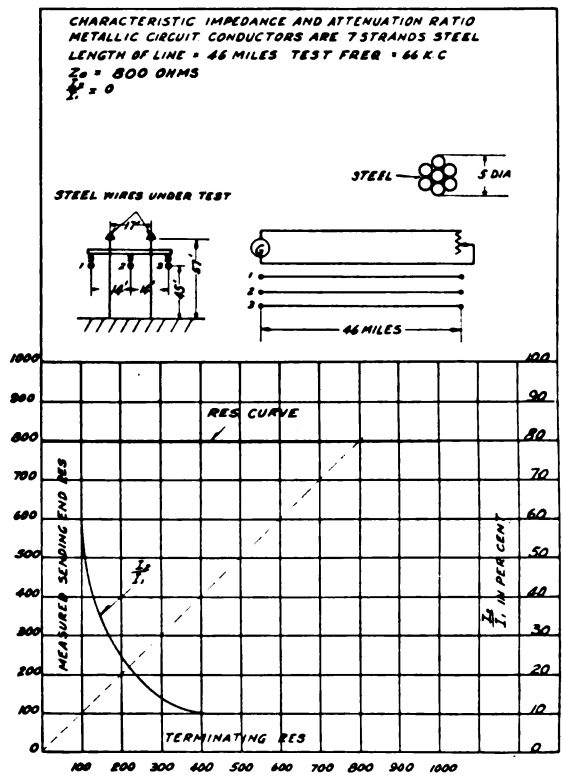

Fig. 26.

values of L and C calculated from line dimensions, we get the velocity of light.

$$V = \frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{14.8 \times 10^{-4} \log \frac{2D}{d} \times \frac{.0194}{\log \frac{2D}{d}} \times 10^{-6}}}$$

$= 186{,}000$ miles per sec.

The assumption that the velocity of propagation on over-

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head conductors is very nearly the velocity of light may be justified by calculating the velocity from the value of the propagation constant. If the velocity were different from the velocity of light, it would be due to losses in the

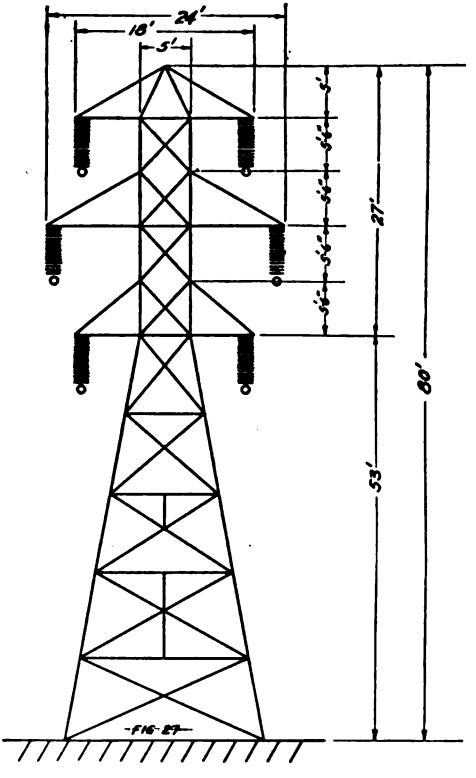

Fig. 27.—Dimensions of tower carrying 132-kv. line on which tests were made.

circuit as represented by the R and G equation. Taking the observed values as

$Z_0 = 850$ ohms $\qquad a = .00923$

$R = 15.7$ ohms $\qquad L = 4.57$ mh.

$G = 0$ $\qquad C = .00633$ $\mu f.$

$$\text{Frequency} = 66{,}000 \text{ cycles}$$

$$P = \sqrt{R + j\omega L} \times \sqrt{G_1 + j\omega C}$$

$$= a + jB$$

$$= \sqrt{15.7 + j\,1890} \times \sqrt{0 + 2.62 \times 10^{-3}}$$

$$= 2.225 \angle 90° - 14.'25$$

$$a = P \cos (90° - 14.'25) = .00923$$

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$B = 2.225$ *radians per mile*

$$\lambda = \frac{2\pi}{B} = \frac{6.2832}{2.225} = 2.83$$

$V = f \times \lambda = 66{,}000 \times 2.83 = 186{,}800$ *miles per sec.*

From this it will be apparent that the velocity of propagation on overhead lines at high frequencies must be very nearly equal to the velocity of light.

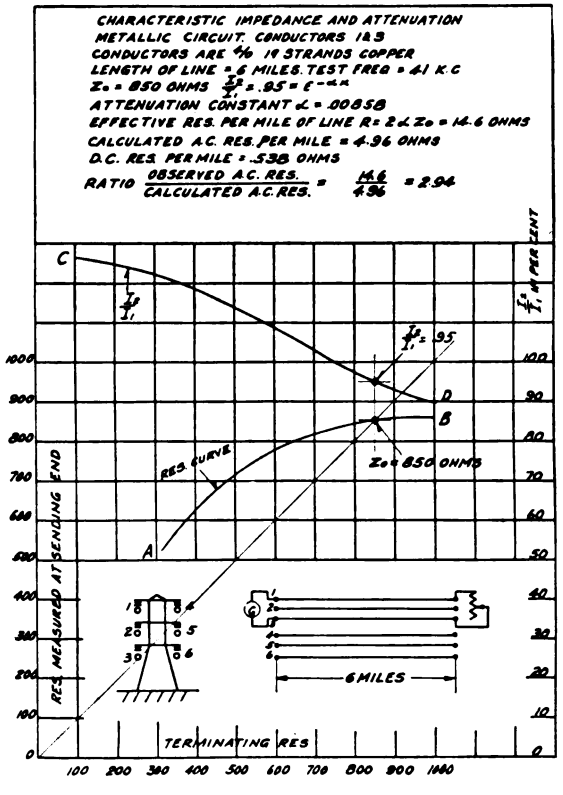

Fig. 28.

By assuming the velocity of propagation to be the velocity of light we are enabled to calculate the effective values of inductance and capacity per mile of line, if Z_0 is accurately determined thus

$$L = \frac{Z_0}{V} \qquad C = \frac{1}{Z_0 V}$$

where V is assumed to be 186,000 miles per second. The values of L and C so determined should not differ greatly

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from those calculated from the dimensions of the system by equations page 60. These relations therefore serve as a check on the accuracy of the observed data.

Tests on Aluminum Steel Conductors

In Fig. 25 is shown data applying to the conductors supported above the power line Fig. 23. These conductors are made of aluminum and steel strands, four strands of steel

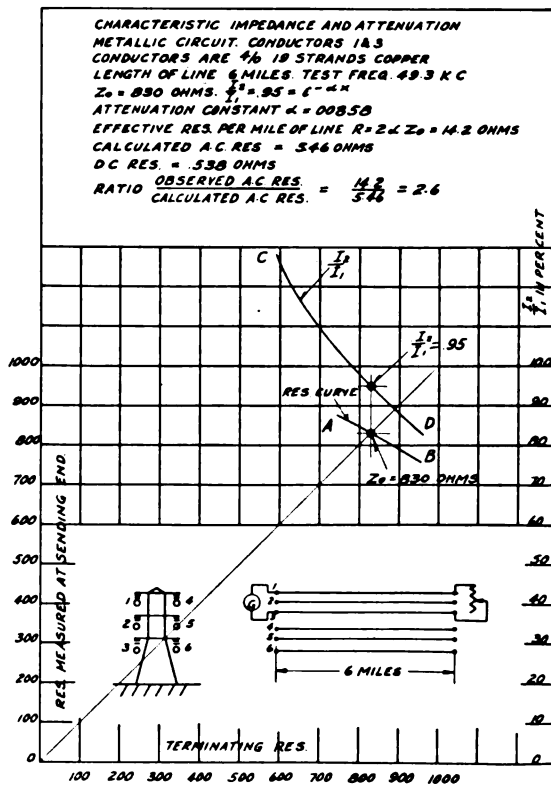
CHARACTERISTIC IMPEDANCE AND ATTENUATION
METALLIC CIRCUIT. CONDUCTORS 1&3
CONDUCTORS ARE 4/0 19 STRANDS COPPER
LENGTH OF LINE 6 MILES. TEST FREQ. 49.3 K C
Zo = 830 OHMS. I2/I1 = .95 = e^-αx
ATTENUATION CONSTANT α = 0085B
EFFECTIVE RES. PER MILE OF LINE R = 2α Zo = 14.2 OHMS
CALCULATED A.C. RES = 3.46 OHMS
D.C. RES. = .538 OHMS
RATIO OBSERVED A.C. RES. / CALCULATED A.C. RES. = 14.2/5.46 = 2.6
I2/I1, IN PERCENT
I2/I1 = 95
RES CURVE
Zo = 830 OHMS
RES. MEASURED AT SENDING END
6 MILES
TERMINATING RES.

Fig. 29.

and three strands of aluminum. The diameter of each strand is .133 inches, the over-all diameter of the conductor being .398. The characteristic resistance is seen from diagram to be 900 ohms. The attenuation ratio $\frac{I_2}{I_1} = .226 = e^{-ax}$. From this we get the attenuation constant $a = .0248$, the distance x being 60 miles. The effective resistance is

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$R = 2a\,Z_0 = 2 \times .0248 \times 900 = 44.5$ *ohms per mile*

The calculated ratio of a-c. to d-c. resistance using Bureau of Standards methods is $\frac{R}{R_0} = 7.22$ assuming the conductor to be 7 strands of aluminum. If we assume that the field distribution in the actual aluminum and steel cable is

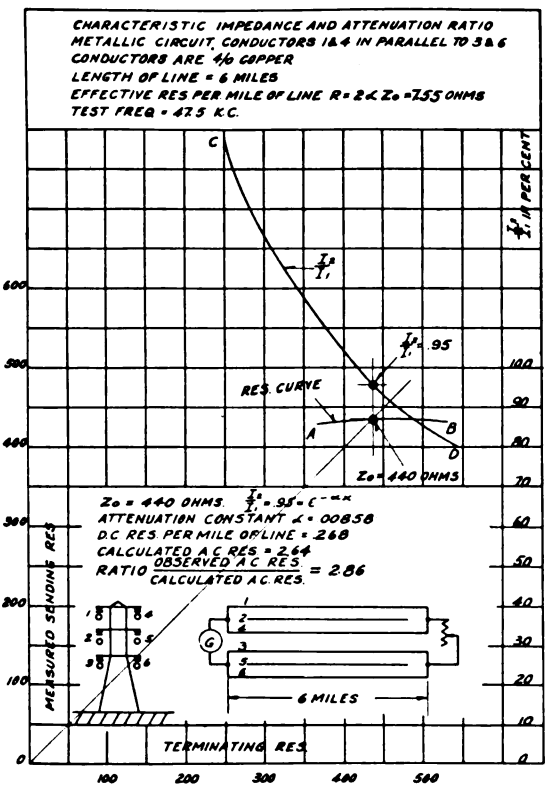

Fig. 30.

substantially the same as if the strands were all aluminum, but using the actual d-c. resistance of the aluminum strands alone, we get $R = 7.22 \times 3.4 = 24.5$ *ohms* as the calculated resistance of one loop mile at 66,000 cycles. The ratio of observed effective resistance to the calculated conductor resistance is

$$\frac{\textit{observed resistance}}{\textit{calculated resistance}} = \frac{44.5}{24.5} = 1.81$$

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Hence there are line losses other than those in the conductors amounting to 81% of the conductor losses.

The calculated value of the characteristic resistance is

$$Z_0 = 276 \log_{10} \frac{2D}{d} = 276 \log_{10} \frac{2 \times 204}{.398} = 832 \text{ ohms}$$

$$\frac{Z_0 \text{ observed}}{Z_0 \text{ calculated}} = 1.08$$

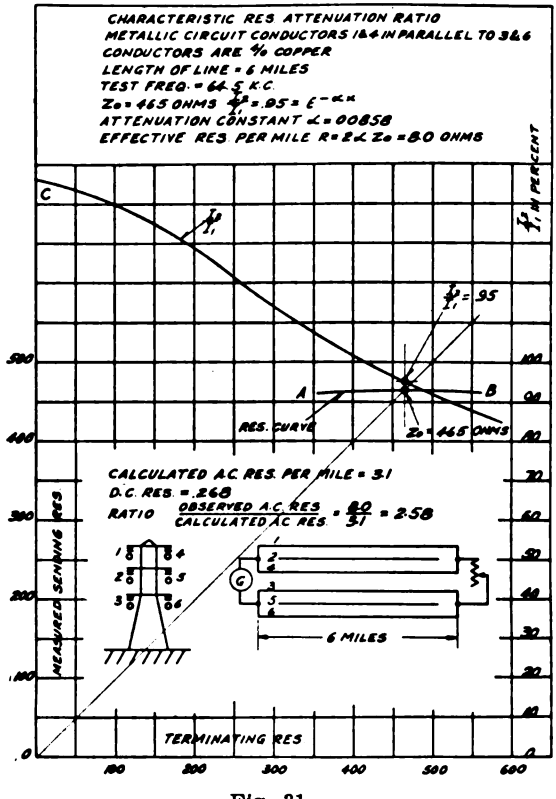

Fig. 31.

Hence the observed value is 8% higher than that obtained by calculation from the physical dimensions of the line.

Test on ½ in. Steel Line

Tests were made on a line consisting of two half-inch steel conductors, each conductor consisting of seven steel strands. The data are plotted in Fig. 26. The attenuation

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on this line was so high that virtually no current was received at the distant end 46 miles away. The characteristic resistance, however, was determined to be 800 ohms. This was simply the resistance of the line as measured at the sending end, the attenuation being so high that there was no observable difference whether the distant end was open or short. This line therefore in its natural condition acted

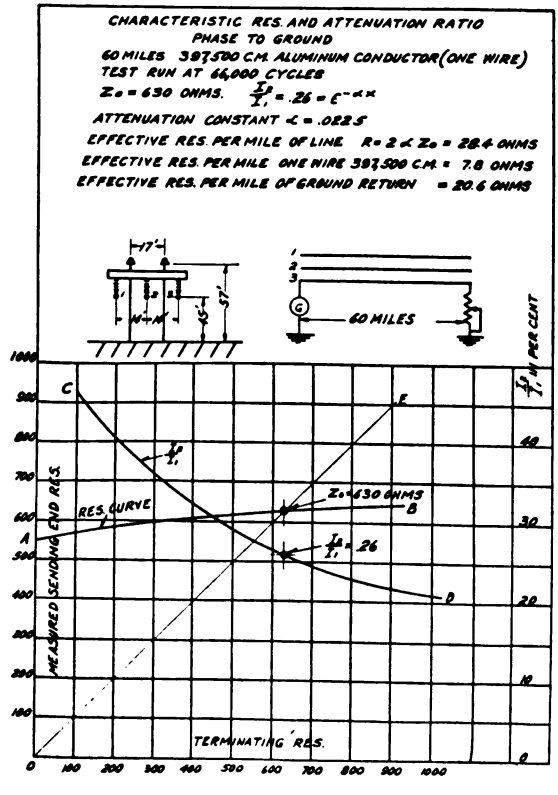

Fig. 32.

like a really infinite line. The value of the characteristic resistance in this case was calculated to be

$$Z_0 = 276 \times \log_{10} \frac{2 \times 204}{.5} = 810 \text{ ohms}$$

The observed value of Z_0 being 800 ohms was very nearly equal to the observed value.

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Tests on 4/0 Copper Double Circuit Line

Tests were made on a double circuit power line built for 132,000 volts. The conductors were No. 4/0 copper, consisting of 19 strands of 105.5 cir. mils diameter. The outside diameter of the conductors was .528 inches. Fig. 27 shows the spacing of the conductors and Fig. 28 shows tests run at 41,000 cycles, using conductors 1 and 3 only. The other

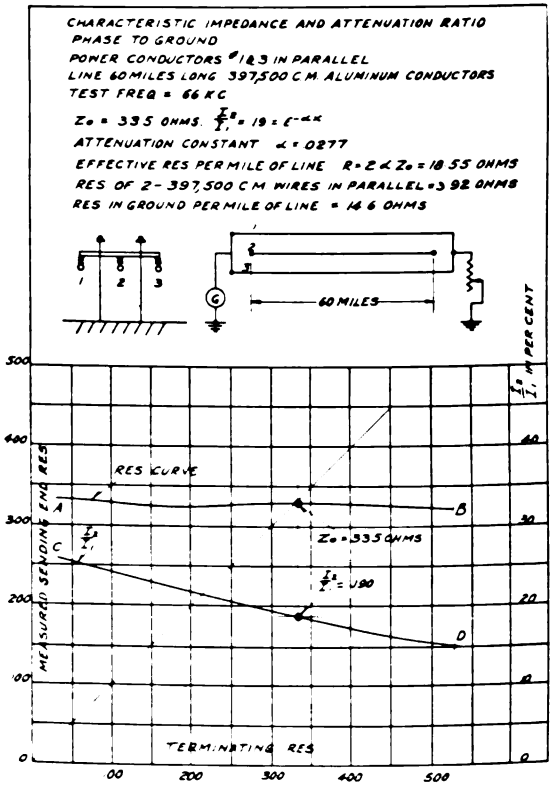

Fig. 33.

conductors were open at both ends. The characteristic resistance is seen to be 850 ohms, while the attenuation ratio is $\frac{I_2}{I_1} = .95 = \epsilon^{-ax}$. The length of the line in this case is only six miles, so $x = 6$ from which we get $a = .00858$, which is the attenuation constant. The resistance of the line per mile is

$R = 2\ a\ Z_0 = 2 \times .00858 \times 850 = 14.6$ *ohms per mile.*

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The ratio of a-c. to d-c. resistance at 41,000 cycles for this conductor is calculated by the method of the Bureau of Standards to be $\frac{R}{R_o} = 9.23$. The d-c. resistance of No. 4/0 copper conductor per loop mile is .538 ohms. Hence the

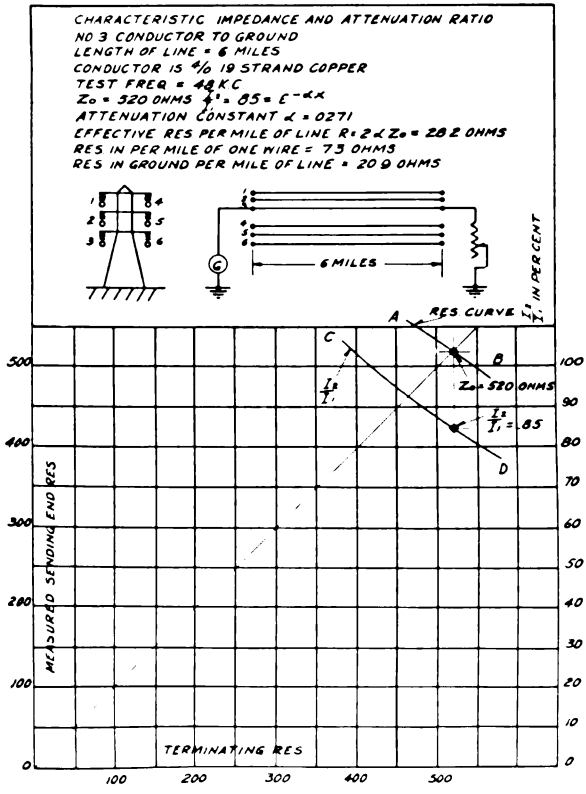

Fig. 34.

calculated a-c. resistance is $9.23 \times .538 = 4.96$ *ohms*.

$$\frac{\text{observed resistance}}{\text{calculated resistance}} = \frac{14.6}{4.96} = 2.94$$

Hence in this circuit it would seem that there must be losses other than those in the copper conductors amounting to 194% of the copper losses.

The spacing of the conductors is 22 feet. This gives a calculated value of characteristic resistance of

$$Z_o = 276 \log_{10} \frac{2D}{d} = 276 \times \frac{\log 2 \times 22 \times 12}{.528} = 828 \text{ ohms}$$

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$$\frac{Z_0 \text{ observed}}{Z_0 \text{ calculated}} = \frac{850}{828} = 102.5\%$$

Similar tests were run on the same conductors at 49.3 kc. and are plotted in Fig. 29. The characteristic resistance $Z_0 = 830$ *ohms*. The attenuation ratio is $\frac{I_2}{I_1} = .95 = e^{-ax}$. The attenuation constant is $a = .00858$ from which

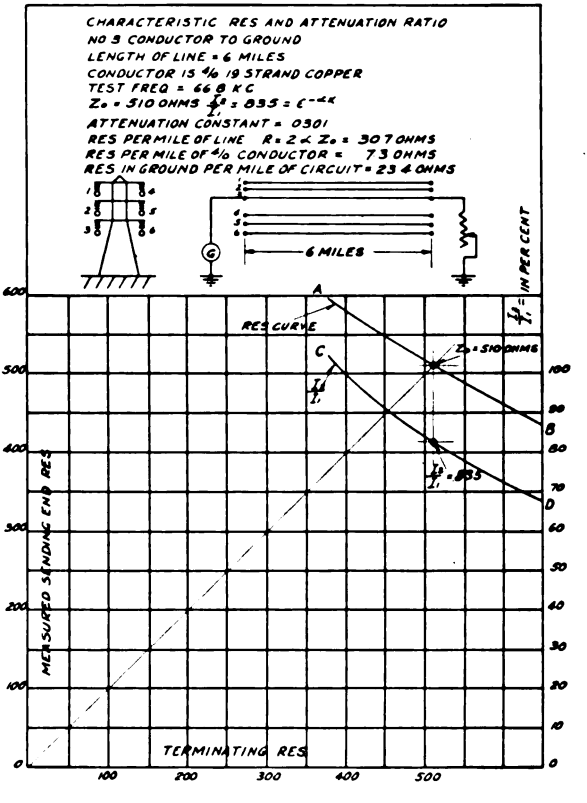
CHARACTERISTIC RES AND ATTENUATION RATIO
NO 3 CONDUCTOR TO GROUND
LENGTH OF LINE = 6 MILES
CONDUCTOR IS 4/0 19 STRAND COPPER
TEST FREQ = 66.8 K.C.
Z₀ = 510 OHMS
ATTENUATION CONSTANT = 0301
RES PER MILE OF LINE R = 2 Z₀ = 30.7 OHMS
RES PER MILE OF 4/0 CONDUCTOR = .73 OHMS
RES IN GROUND PER MILE OF CIRCUIT = 23.4 OHMS
6 MILES
RES CURVE
MEASURED SENDING END RES.
TERMINATING RES.
I₂/I₁ IN PER CENT

Fig. 35.

the effective resistance is 14.2 ohms. The calculated value of $\frac{R}{R_0} = 10.15$ which gives the calculated conductor resistance as 5.46 ohms.

$$\frac{\text{observed resistance}}{\text{calculated resistance}} = \frac{14.2}{5.46} = 2.6$$

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Tests on Two 4/0 Copper Lines in Parallel

In Fig. 30 are shown data for a circuit consisting of wires 1 and 4 in parallel as one side of the circuit, and wires 3 and 6 in parallel as the other side or return. The characteristic impedance $Z_0 = 440$ *ohms*. The attenuation ratio is $\frac{I_2}{I_1} = .95$. The circuit is the same as shown in Fig. 27 supported on steel towers with 4/0 copper conductors and is six miles long. The attenuation constant is .00858. The effective resistance per mile of circuit is

$$R = 2a\,Z_0 = 2 \times .00858 \times 440 = 7.55 \text{ ohms}$$

The resistance of one of the circuits alone would be

$$2 \times 7.55 = 15.1 \text{ ohms per mile of single circuit.}$$

This corresponds to the previous data giving 14.6 ohms per mile. The test was made at 47,500 cycles. The data is not sufficiently accurate to draw exact conclusions as regards the resistance of wires in parallel. A very slight error in the value of Z_0 makes an appreciable change in the attenuation ratio on so short a line, and therefore gives a different value of calculated resistance.

Fig. 31 gives similar data on the same line except that the test was made at 64,500 cycles. The characteristic resistance is 465 ohms. The attenuation ratio $\frac{I_2}{I_1} = .95$. The attenuation constant is as before .00858. The value of circuit resistance per mile is

$$R = 2a\,Z_0 = 2 \times .00858 \times 465 = 8.0 \text{ ohms.}$$

This makes the resistance of each circuit by itself $2 \times 8 = 16$ *ohms*, which is in general agreement with the previous data.

From these data it is evident that the resistance of one mile of 4/0 copper conductor lies between 14.6 and 16.0 ohms per loop mile depending upon the frequency. The resistance per mile of one wire is therefore 7.3 ohms to 8.0 ohms, and we may use these values in future calculations with confidence.

Ground Return Circuit

There has been considerable discussion regarding the relative merits of a line having complete metallic circuit and a line with ground return. In making measurements on

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ground return circuits employing the usual methods, serious errors have been introduced, and wrong conclusions drawn. Even with reasonable precautions using improved methods of measurement, erroneous conclusions are likely to be drawn from test data unless due allowances are made for the resistances of the ground stake at both the sending and receiving end. In most test data on ground return cir-

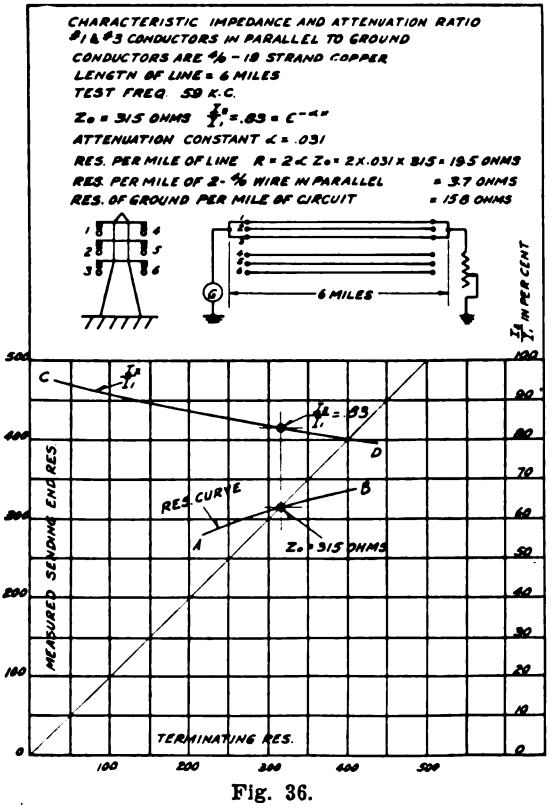

Fig. 36.

cuits the effect of ground stake resistance has been entirely neglected. By taking these details into account and calculating the circuit resistances according to the preceding method, considerable interesting information is brought to light.

One Conductor to Ground

In Fig. 32 are shown the data on a sixty-mile circuit having ground return. The conductor is 397,500-cir. mil.

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aluminum cable. The measurements plotted in the figure apply to the case where only one conductor was used, the remaining conductors being insulated at both ends. The line construction is the same as shown in Fig. 23. The test was run at 66 kilocycles. The curves as plotted have been corrected for ground stake resistance at both ends. The

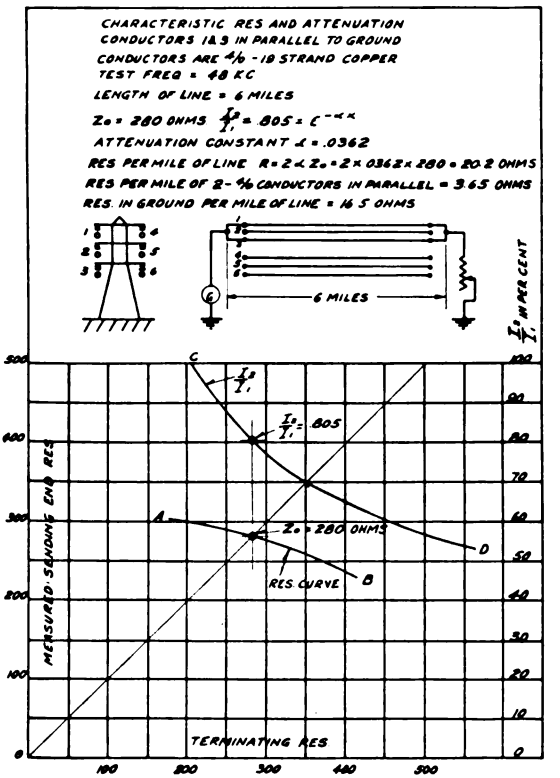
CHARACTERISTIC RES AND ATTENUATION
CONDUCTORS 1 & 3 IN PARALLEL TO GROUND
CONDUCTORS ARE 4/0 - 19 STRAND COPPER
TEST FREQ = 66 KC
LENGTH OF LINE = 6 MILES
Z₀ = 280 OHMS
ATTENUATION CONSTANT
RES PER MILE OF LINE
RES PER MILE OF 2 - 4/0 CONDUCTORS IN PARALLEL = 3.65 OHMS
RES. IN GROUND PER MILE OF LINE = 16.5 OHMS
6 MILES
MEASURED SENDING END RES.
TERMINATING RES
RES. CURVE
Z₀ = 280 OHMS
.805
A
B
C
D

Fig. 37.

characteristic resistance is found to be $Z_0 = 630$ *ohms*. The attenuation ratio $\frac{I_2}{I_1} = .26 = e^{-ax}$ from which we get the attenuation constant $a = .0225$, the length of the line x being 60 miles. The effective resistance per mile of circuit is

$$R = 2aZ_0 = 2 \times .0225 \times 630 = 28.4 \text{ ohms}$$

This is the total resistance per mile including all losses. The resistance of the cable used in this test was previously determined as in Fig. 22 to be 15.7 ohms per loop mile. The

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resistance per mile of one wire is $\frac{15.7}{2} = 7.85$ *ohms*. Subtracting this from the total resistance per mile of circuit we get 20.6 ohms as the resistance per mile, excluding the resistance in the metallic conductor, and also the dielectric and radiation losses associated with it. This 20.6 ohm resistance per mile must therefore *represent the resistance per mile of ground return path.*

The existence of a definite measurable resistance in the ground per mile of line is in striking contrast to the ideas commonly held regarding ground resistance in general. It is generally believed that the resistance of the ground part of a ground return system is so low as to be negligible in comparison with the conductor resistance of the line. The only resistances regarded as of practical importance are those occurring at the ground stakes or at whatever point the circuit makes contact with the ground. For direct current and for 25 or 60 cycles this view is admitted to be correct. The low ground resistance in the case of d-c. or power frequencies is due to the ground currents spreading out over enormous areas after the current once really enters the ground.

At high frequencies such as those under which the accompanying tests were made, the ground resistance seems to be very appreciable and to have a rather definite value. The fact that the ground does present a resistance of approximately 20 ohms per mile of circuit, indicates that the ground current does not spread out over a very wide area. If we assume that the ground resistance per foot cube is 250 ohms, then the cross sectional area of the ground path which gives a resistance of 20 ohms per mile is 66,000 square feet. That is, if the current were uniformly distributed under the line over a strip 1000 feet wide and a depth of 66 feet, the resistance would be 20 ohms per mile. It is not intended to convey the idea that the current is uniformly distributed, but merely to suggest the order of magnitude of the cross sectional area of the ground return path. The actual current density is greatest at the surface and immediately under the line. The density falls off rapidly with the depth and somewhat less rapidly with the width. The analysis of current distribution in the ground is a problem similar to skin effect on conducting wires, the density being greatest at the surface.

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Two Conductors in Parallel to Ground

In Fig. 33 is given results of a test made on the same line as in Fig. 32 except that the two outside conductors are connected in parallel. The characteristic impedance $Z_0 = 335$ *ohms*. The attenuation ratio is $\frac{I_2}{I_1} = .19 = e^{-ax}$ which gives

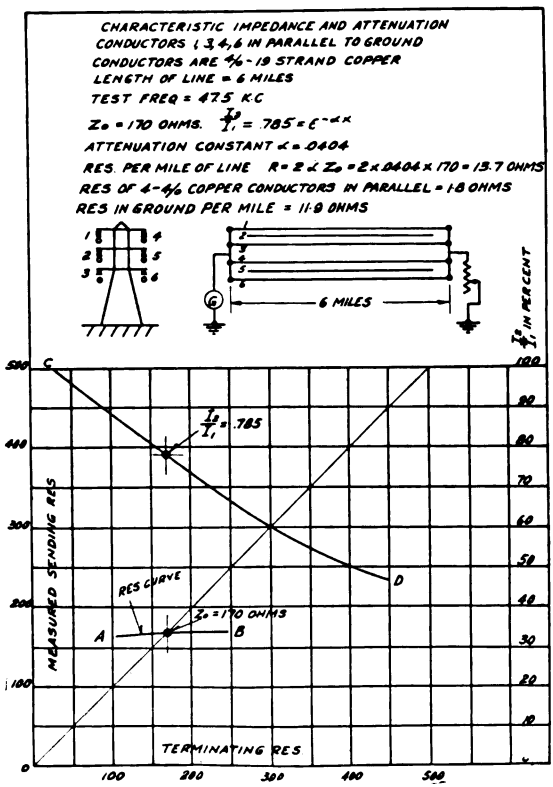

Fig. 38.

the attenuation constant $a = .0277$. The effective resistance per mile of circuit is

$$R = 2aZ_0 = 2 \times .0277 \times 335 = 18.5 \text{ ohms}$$

The resistance of two 397,500-cir. mil aluminum conductors in parallel is 3.92 ohms. Subtracting this from the total of 18.5 ohms we get 14.6 ohms as the resistance chargeable to the ground return per mile of circuit.

It is of considerable interest to note that the resistance

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of the ground return path is less when two conductors are used in parallel, than with a single conductor.

Tests on Double Circuit 4/0 Copper Line One Wire to Ground

In Fig. 34 are given the results of a test made on one conductor of a double circuit line shown in Fig. 27. The conductors are 4/0 19 strands copper. The length of the line is 6 miles. One of the bottom conductors indicated as No. 3 was connected to ground, while all other conductors were open at both ends. The test is similar to that shown in Fig. 32. The test frequency was 48 kc. The characteristic resistance $Z_0 = 520$ *ohms*. The attenuation ratio is $\frac{I_2}{I_1} = .85 = e^{-ax}$ which gives the attenuation constant $a = .0271$. The effective resistance per mile of line is

$$R = 2aZ_0 = 2 \times .0271 \times 520 = 28.2 \text{ ohms}$$

The resistance per mile of one 4/0 copper conductor is 7.3 ohms which leaves 20.9 ohms as the resistance chargeable to the ground return.

Fig. 35 is a test similar to Fig. 34 except that the test frequency is 66.8 kilocycles. The value of Z_0 is 510 ohms and the attenuation ratio $\frac{I_2}{I_1} = .835 = e^{-ax}$. The attenuation constant a is .0301 and the resistance per mile of line is

$$R = 2aZ_0 = 2 \times .0301 \times 510 = 30.7 \text{ ohms}$$

Subtracting 7.3 ohms for the copper conductor we have 23.4 ohms as the resistance per mile of ground return.

Two Conductors in Parallel to Ground

In Fig. 36 are given test data on the same line as Fig. 35 except that conductors number 1 and 3 are connected in parallel. The frequency is 59 kc. The value of Z_0 is 315 ohms. The attenuation ratio $\frac{I_2}{I_1} = .83$. The attenuation constant $a = .031$. The resistance per mile of line is

$$R = 2aZ_0 = 2 \times .031 \times 315 = 19.5 \text{ ohms}$$

Subtracting 3.7 ohms, the resistance of two 4/0 wires in parallel leaves 15.8 ohms as the resistance per mile of ground return. This result compares with 14.6 ohms in Fig. 33.

In Fig. 37 are given the results of a test similar to Fig.

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36, except that the test frequency is 48 kc. The value of Z_0 is 280 ohms. The attenuation ratio is $\frac{I_2}{I_1} = .805$. The attenuation constant is .0362. The resistance of the circuit per mile is

$$R = 2aZ_0 = 2 \times .0362 \times 280 = 20.2 \text{ ohms}$$

Subtracting 3.65 ohms as the resistance of two conductors

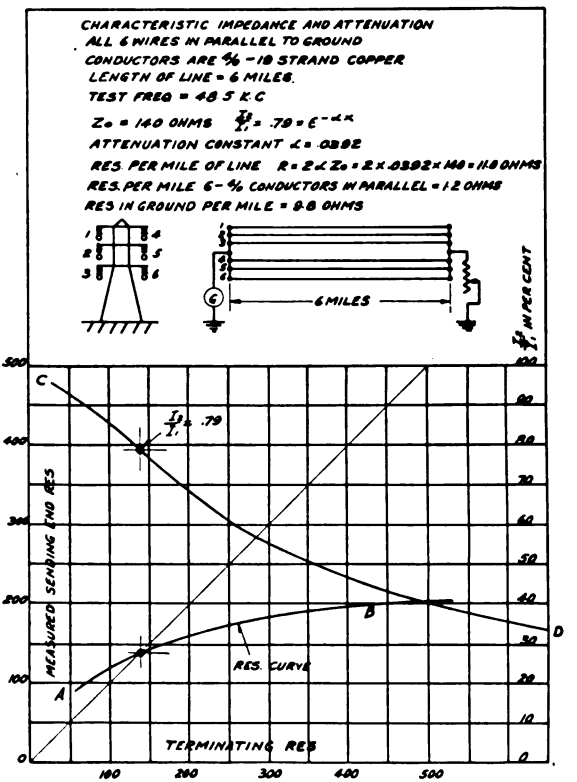

Fig. 39.

in parallel we get 16.5 ohms as the resistance per mile of ground return. This compares with the value of 15.8 ohms in Fig. 36.

Four Conductors in Parallel to Ground

In Fig. 38 are given the test data on the same line as in Fig. 37, except that conductors 1, 3, 4 and 6 are tied in parallel to ground. The test frequency is 47.5 kc. The

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characteristic impedance $Z_0 = 170$ *ohms*. The attenuation ratio $\frac{I_2}{I_1} = .785 = e^{-ax}$. The attenuation constant $a = .0404$. The resistance per mile of line is

$$R = 2aZ_0 = 2 \times .0404 \times 170 = 13.7$$

The resistance of four 4/0 conductors in parallel is 1.8 ohms. Subtracting this from the total gives 11.9 ohms per mile as the resistance of the ground return.

Six Conductors in Parallel to Ground

In Fig. 39 are given the results of a test similar to Fig. 31, except that all six conductors are tied in parallel to ground. The value of the characteristic impedance is $Z_0 = 140$ *ohms*. The attenuation ratio $\frac{I_2}{I_1} = .79 = e^{-ax}$. The attenuation constant $a = .0392$. The resistance per mile of line is

$$R = 2aZ_0 = 2 \times .0392 \times 140 = 11.0 \text{ ohms}$$

The resistance of six 4/0 copper conductors in parallel is 1.2 ohms. Subtracting this from the total gives 9.8 ohms as the resistance per mile of ground return.

Discussion of Results

It will be noticed that the ground resistance indicated by the preceding tests is progressively less each time conductors are added in parallel. The explanation of this is not clear at present as it would seem that the ground resistance should remain substantially constant.

In making these tests it will be remembered that all wires not included in the tests were open at both ends. It has been suggested that currents are induced in these idle conductors and that the losses in these appear as added resistance in the circuit under test. Such losses of course are not chargeable against ground resistance.

This explanation is not tenable since it can be shown that no currents flow in a wire paralleling an energized wire provided it is insulated throughout and open at both ends.

Although the interpretation of the results applying to ground return circuits may be open to question so far as the actual magnitude of the ground resistance is concerned, the

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data would seem to establish beyond question the fact that the ground does have a very definite resistance per mile and that this resistance is comparable with the resistance of the conductors. It is this ground resistance which makes the attenuation of a ground return circuit higher than for a metallic circuit. This attenuation, however, is not nearly so great as is generally supposed.

Studies on power lines at high frequencies such as the foregoing are of value entirely aside from the question of communication. For example, the values of characteristic impedance, attenuation in metallic and in ground return circuits have immediate application in the study of many problems which arise from switching surges and lightning protection. The methods here developed in the analysis of antenna coupling may be used to calculate the reduction of induced potential due to the presence of ground wires above a power line.

The reactions taking place in the ground portion of a ground return circuit are similar to those taking place on the ground side of an advancing radio wave. A further study of ground return circuits may throw considerable light on radio phenomena.

SUMMARY

Communication is a matter of vital importance in the operation of large power systems. The difference between ordinary wire communication and communication over high voltage power lines is discussed. The necessity for using energy levels greatly in excess of that commonly used on ordinary wire lines is indicated. The importance of selecting and utilizing only those circuit elements which are inherently stable is discussed. The superior stability of the two frequency system of duplex is brought out. A standard 250-watt communication equipment is described in detail. The great value of the heterodyne principle for calling is noted. The theory of antenna coupling is developed beginning with elementary principles. Several numerical examples are given to show the application of the theory. It is shown that the efficiency of antenna coupling is quite high contrary to the general opinion prevailing at present. A brief description of the propagation of high frequency

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currents over wires together with some of the principal equations is given. Tests are given which show a close agreement with accepted line theory. A simple method of determining line characteristic resistance and method of determining the effective line resistance is described. This method is applied to ground return circuits and discloses the existence of considerable resistance per mile in the ground return path.

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MEASUREMENTS OF RADIO FREQUENCY AMPLIFICATION

By
Sylvan Harris

The need for a simple and accurate method of measuring radio frequency amplification and for studying the over-all characteristics of radio receivers is urgent, and it is hoped that the work described in this paper may help somewhat to fill this need. Many investigators have made measurements of "gain-per-stage" of radio frequency amplifiers, and in all cases which have come to the attention of the writer, such measurements required a knowledge of the values of input and output voltages.

Due to the difficulty of measuring voltages of the order of a few microvolts, the voltage impressed on the state under consideration in these various measurements, was generally far above the value of the radio frequency voltages encountered in radio receivers. In one instance,[1] the impressed voltage was obtained as an inductive drop in a short straight wire surrounded by a concentric return conductor. The voltage was computed from the current flowing in the wire and the inductance of this wire. Although measurements made by this method agreed, as to order of magnitude, with computed values, there is no means of determining the accuracy of the method. Furthermore, this method was applied to a radio frequency amplifier in which the gain-per-stage was great, so that small errors would not be noticeable.

Some experimenters have been in the habit of attempting to measure the gain in an isolated stage, that is, a stage not connected to other stages, as is the case in radio receivers. Such measurements do not give any fair idea of the actual gain, for, due to reaction of the other stages, it may be much different from the values so obtained. Especially is this true

[1] A. W. Hull, Phys. Rev. Vol. 27, No. 4, page 442.

*Received by the Editor April 5, 1927.

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of amplifiers in which there is regeneration present, even when totally shielded.

The advantages of the method described in this paper may be summarized as follows:

(a) the measurements are independent of the values of the input and output voltages;

(b) the measurements are made on the stage in question while actually in a radio receiver, under actual operating conditions;

(c) no connections are made to the stage in question for purposes of measurement, excepting a switch for cutting this stage in and out of the amplifier;

(d) no special apparatus is required for making the measurements other than that usually found in radio laboratories.

The method may be explained as follows: Consider a completely shielded radio receiver having three stages in the radio frequency amplifier, a detector, and an audio frequency amplifier. Then let

K_1 be the gain in the first radio-frequency stage
K_2 be the gain in the second radio-frequency stage
K_3 be the gain in the third radio-frequency stage
K_d be the detector constant
K_a be the gain in the audio-frequency amplifier
e be the alternating voltage impressed at the input terminals of the receiver.
V_1 be the voltage output of the receiver when all the stages are used.
V_2 be the voltage output of the receiver when the second stage is omitted.

When all the stages are connected in, assuming the square law of the detector, the output voltage of the receiver is

$$V_1 = (e K_1 K_2 K_3)^2 K_d K_a \qquad (1)$$

When the second stage of the radio frequency amplifier is omitted, this becomes

$$V_2 = (e K_1 K_3)^2 K_d K_a \qquad (2)$$

The ratio of these two expressions gives, as the gain in the second r. f. stage

$$K_2 = \sqrt{\frac{V_1}{V_2}} \qquad (3)$$

In order to obtain values for V_1 and V_2, a non-inductive

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voltage-divider of about 20,000 ohms was connected to the output of the receiver, as shown in Figure 1. The remainder of the circuit of Figure 1 shows a vacuum tube voltmeter having in its output circuit an audio transformer, carborundum crystal rectifier, and a microammeter. This arrangement was adopted on account of its great sensitivity and convenience in operating.

A "datum" indication of the microammeter is selected, as will be explained later, all measurements being taken with this same deflection. Let this deflection correspond to a voltage input to the vacuum tube voltmeter of v_o, as indicated in Figure 1. The output voltage of the receiver will then be

$$v_o \frac{R}{r} \qquad (4)$$

in which r is that portion of R, across which the voltage drop must be taken in order that the microammeter regis-

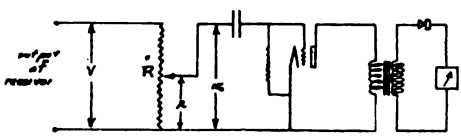

Figure 1

ter the "datum" deflection. By using this arrangement the measurements are made independent of the characteristics of the vacuum tube voltmeter, and associated apparatus, since a completely modulated radio frequency supply is used, the modulation frequency being constant.

Letting r_1 be the value of r when the stage under question is connected in, and r_2 the value when the stage in question is cut out, equation (3) becomes:

$$K_2 = \sqrt{\frac{r_2}{r_1}} \qquad (5)$$

In making the measurements a low power driver was used which supplied radio frequency power completely modulated at 60 cycles per second. The measurements were made in a sheet-iron booth, and the receiver itself was completely shielded, over-all and between stages. Only one wire was led into the booth from the river, the sheet-iron which

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formed the booth acting as the return circuit. The driver was located outside the booth, and delivered power to a small coil located in a copper box fastened on the outside of the booth. This coil acted inductively on another small coil, and between the two coils was a sliding copper "door"; by sliding this door the amount of power delivered by the coil connected to the driver to the other coil, connected to the receiver input, was regulated. In making the measurements care must be taken that the detector is not "overloaded."

The "datum" indication is obtained by observing the smallest deflection of the microammeter that can be read accurately when the weakest signal voltage is impressed on the input of the receiver with the stage in question cut out, and the voltage divider set at maximum. To measure the gain of that stage, it is then switched into the receiver, and the voltage divider adjusted until the datum deflection is

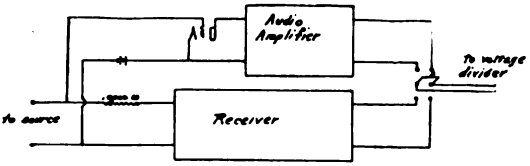

Figure 2

obtained again. Eq. (5) then gives the gain in that stage (where $r_2=R$).

In plotting curves of gain against frequency the datum deflection that is used for each measurement is unimportant; the only requirement is that the detector stage be not overloaded. In order to obtain a check on the method and at the same time determine when overloading of the detector occurs, another vacuum tube voltmeter, followed by an audio frequency amplifier, was connected to the input of the receiver, as shown in Figure 2. On account of the fact that the input voltage required at the set is so small and it is difficult to obtain an amplification in this amplifier of the same order of magnitude as that which the receiver furnishes, a resistance of 10,000 ohms was interposed between the receiver input and the points at which the second vacuum tube voltmeter was connected. In later measurements this resistance was replaced by a small condenser, simulating the capacity of an antenna.

This vacuum tube voltmeter and amplifier serves as a

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device for measuring the input voltage. The double-throw switch enables the voltage divider to be connected either to the input measuring system or to the output of the receiver. Assuming a square law in the receiver, and a square law vacuum tube voltmeter at the input, the relation between the settings of the voltage divider when the switch is thrown first in one direction and then in the other, should be linear. Or, using the subscript i to indicate input measure-

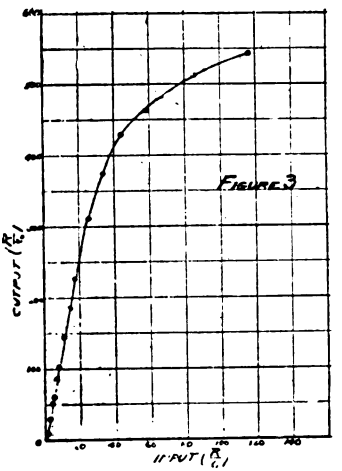

Figure 3

ments and apparatus, and the subscript o for the output measurements, other symbols remaining the same:

$$\frac{R}{r_1} v_o = e^2 K_{d1} K_{a1}$$

$$\text{or} \qquad e = \sqrt{\frac{\frac{R}{r_1} v_o}{K_{d1} K_{a1}}} \tag{6}$$

Likewise,

$$\frac{R}{r_o} v_o = (e K_1 K_2 K_3)^2 K_d K_a$$

$$\text{or} \qquad e = \sqrt{\frac{\frac{R}{r_o} {}^{o}a}{(K_1 K_2 K_3)^2 . K_d K_a}} \tag{7}$$

Equating these and reducing:

$$r_o = a \; constant \; x \; r_1 \tag{8}$$

Figure 3 illustrates this relation; it is a curve obtained

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at constant frequency, the input voltage varying. It is linear for voltages up to the point of overload of the detector in the receiver. The same curve has been plotted in Figure 4, this time taking the square-root of the "input" values. Plotted to logarithmic coordinates, this curve is likewise linear up to the point of overload of the detector, and has a slope very close to 2, as would be expected from the law of the detector. Figure 4 then, is the response curve of the receiver at constant frequency, varying input voltage. Figure 5 is the

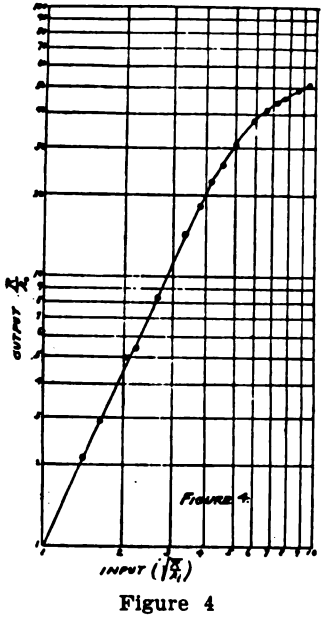

Figure 4

gain-frequency curve of a stage in an experimental receiver.

The same set-up can be used to study the over-all characteristics of radio receivers, at various frequencies, and at constant input voltage. The vacuum tube voltmeter is then used only to indicate the constancy of the input voltage. The over-all gain of the receiver is given by

$$K_T = \frac{\frac{R}{r_o} v_o}{\sqrt{\frac{\frac{K}{r_1} v_o}{K_{d1} K_{a1}}}} = a\ constant\ x\ \frac{\sqrt{r_1}}{r_o} \qquad (6)$$

Keeping the input voltage constant, represented by r_1 in eq.

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(9), the amplification of the receiver is inversely proportional to r_o. Such a curve of an experimental receiver is shown in Figure 6.

This method has also been used by the writer for determining the detection coefficients of detectors. In measurements of this kind, as also in obtaining curves like that in

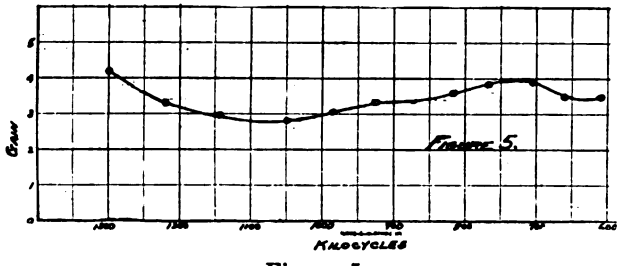

Figure 5

Figures 3, 4 and 6, the actual values obtained are arbitrary, but with a careful set-up conditions can easily be reproduced and the same datum used each time a set of measurements is taken. All the measurements will then be reduced to a common basis, affording an easy means of comparison.

If it is desired the vacuum tube voltmeter at the input may be calibrated, and the amplification of the audio fre-

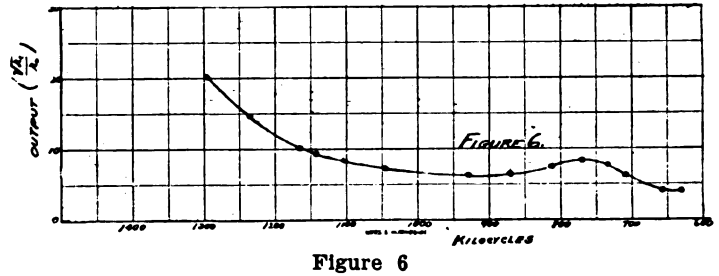

Figure 6

quency amplifiers can be determined. But this will lead to difficulties which this method seeks to avoid.

In conclusion, it may be stated that the only serious difficulty in making these measurements is the avoidance of a "noise level" which, in the present work, was sometimes greater than the original signal strength used to determine the "datum". The noise level was mainly due to vibrations of the building. Its effect on the accuracy of the measure-

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ments was appreciable only when using the weakest input voltages. In figure 3, for example, it was difficult to obtain points on the curve below $R/r_1 =$ about 20 in the daytime, so much of the work was done at night when the shaking of the building was the least.

SUMMARY—A method of measuring radio frequency amplification in radio receivers is presented. It applies accurately only to non-regenerative receivers, to the receivers in which regeneration is not very pronounced. The set-up described in the paper can be used without change for studying the gain-per-stage, the over-all characteristics of a receiver, and detection co-efficients of electron tubes. The method has the advantage that the measurements do not require a knowledge of the actual values of the voltages, and do not require the removal of a stage in question from the receiver. The measurements are made with the stage or the complete receiver under actual operating conditions.

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BOOK REVIEW

Alternating Current Rectification and Allied Problems, by L. B. W. JOLLEY. JOHN WILEY & SONS, INC., SECOND EDITION. 472 + XXII Pages, 340 illustrations. Price $6.

This volume, of English origin, gives a nearly complete resume of European systems of alternating current rectification and includes details of many American appliances as well. Besides the commonly known types (such as rotary and vibrating mechanical rectifiers, mercury vapor systems, thermionic devices and gas filled tubes) many of the lesser known devices are described as well. To any one who does not follow current literature of European countries many of the latter may appear new and several would seem to offer much promise for further investigation.

The book is hardly complete in the matter of recent American practice, since several of the popular radio rectifiers have been neglected, such as the full wave filamentless tubes, cuprous oxide rectifier plates, magnesium plate rectifiers and the like. The date of the second edition preface (June, 1926) follows the introduction of these devices in this country by a considerable time. On the other hand some details of a type of paste rectifier recently popularized here, following the discoveries of Andre', (as a cartridge type of rectifier for trickle charger work) are included.

In this age of super-power, alternating current has forged ahead, mainly due to the ease of obtaining the necessary high voltages, in spite of many disadvantages that are inherent with the a. c. system. While an outline of the most plausible methods of rectification with the ultimate view of converting large amounts of power to potentials of hundreds of kilovolts is of interest primarily to the power engineer, the methods are of interest to the radio engineer as well. Indeed, it seems that the majority of the systems originated as a solution to radio problems. All types described have been used by radio engineers.

The book is written in a popular style, although the general description of each system and the detailed account of

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the outstanding problems are followed by a mathematical study of the theory. In general the formulae given are complete and final and do not involve mathematics higher than algebra and trigonometry. An excellent bibliography appears at the end of each section, comprising some seven hundred items in all, from over seventy periodicals as well as a number of text books. Whenever a particular paper listed in these bibliographies also appears in "Science Abstracts" the file reference to this periodical is also given. The reader might be saved a little time in looking up these references if the tabulation of the particular subject contained therein were a little more specific.

A special section has been included in this edition, to the listing of the sources of power for broadcast receivers and an analysis of the advantages and disadvantages of all methods are discussed. Other special sections include harmonic analysis, wave form measurements, inverters and the use of rectifiers in the field of measurements.

"Alternating Current Rectification" is an excellent reference book and should prove a valuable addition to the library of any radio engineer.

R. R. Batcher.

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DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued April 5, 1927—May 17, 1927

By
JOHN B. BRADY

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,623,152—RADIO APPARATUS—A. M. YOUNG, of Brentwood, and N. W. SIMPSON, of Bolton, England. Filed June 12, 1922, issued Apr. 5, 1927. Assigned to Radio Corp. of America.

1,623,360—CONDENSER—T. T. RODGERS, M. A. GIBLIN, F. K. MOORE, and G. T. MOORE, of Milwaukee, Wis. Filed Sept. 24, 1924, issued Apr. 5, 1927. Assigned to Radio Corp. of America.

1,623,563—ELECTRICAL CONDENSER—G. F. A. STONE, London, England. Filed July 15, 1925, issued Apr. 5, 1927.

1,623,628—COMBINED A BATTERY TRICKLE CHARGER AND BY SUPPLY—D. R. LOVEJOY, of New York City. Filed Oct. 16, 1926, issued Apr. 5, 1927. Assigned to Lovejoy Development Corp.

1,623,741—VARIABLE ELECTRICAL CONDENSER—L. R. McDONALD, Westmount, Quebec, Canada. Filed Nov. 10, 1922, issued Apr. 5, 1927.

1,623,742—VARIABLE ELECTRICAL CONDENSER—L. R. McDONALD, Westmount, Quebec, Canada. Filed Mar. 7, 1924, issued Apr. 5, 1927.

1,623,745—TRANSMITTING SYSTEM FOR RADIANT ENERGY—A. F. MURRAY, Boston, Mass. Filed Jan. 14, 1921, issued Apr. 5, 1927. Assigned to John Hays Hammond.

1,623,918—RADIO RECEIVING APPARATUS—A. H. GREBE, Hollis, N. Y. Filed Dec. 17, 1924, issued Apr. 5, 1927.

1,623,966—RADIO TRANSMISSION SYSTEM—P. S. CARTER, New Brunswick, N. J. Filed June 25, 1923, issued Apr. 12, 1927. Assigned to Radio Corp. of America.

1,623,005—METHOD OF RADIO SIGNALING—M. OSNOS and L. KASARNOWSKY, Berlin, Germany. Filed Jan. 12, 1923, issued Apr. 12, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie.

1,624,006—CIRCUIT ARRANGEMENT FOR HIGH FREQUENCY SENDING STATIONS—M. OSNOS, Berlin, Germany. Filed Feb. 12, 1923, issued Apr. 12, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie.

1,624,148—SIGNAL RECEIVING CIRCUITS—E. L. POWELL, Washington, D. C. Filed Mar. 6, 1926, issued Apr. 12, 1927. Assigned to Wired Radio, Inc.

1,624,185—WIRELESS RECEIVING CIRCUITS—H. J. ROUND, London, England. Filed May 21, 1923, issued April 12, 1927. Assigned to Radio Corp. of America.

1,624,201—RADIO SIGNALING—C. S. AGATE and P. W. WILLANS, Chelmsford, and London, England, respectively. Filed Sept. 28, 1921, issued Apr. 12, 1927.

1,624,208—HIGH FREQUENCY SIGNALING—P. M. J. BOUCHEROT, Paris, France. Filed Aug. 29, 1921, issued Apr. 12, 1927.

1,624,215—RADIO SIGNALING SYSTEM—G. H. CLARK, Brooklyn, N. Y. Filed June 27, 1922, issued Apr. 12, 1927. Assigned to Radio Corp. of America.

1,624,332—ELECTRICAL CONDENSER—P. E. GILLING, East Orange, N. J. Filed Feb. 23, 1925, issued Apr. 12, 1927.

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1,624,333—ELECTRICAL CONDENSERS—P. E. GILLING, East Orange, N. J. Filed Mar. 13, 1925, issued Apr. 12, 1927.

1,624,334—ELECTRICAL CONDENSER—P. E. GILLING, East Orange, N. J. Filed Mar. 13, 1925, issued Apr. 12, 1927.

1,624,451—VACUUM TUBE—H. W. WEINHART, Elizabeth, N. J. Filed Apr. 9, 1921, issued Apr. 12, 1927. Assigned to Western Electric Co.

1,624,459—SHIELDING AND BALANCING MEANS—W. J. ADAMS and A. HADDOCK, of Irvington, N. Y., and East Orange, N. J. respectively. Filed Dec. 20, 1920, issued Apr. 12, 1927. Assigned to Western Electric Co.

1,624,473—HIGH FREQUENCY SIGNALING SYSTEM—L. M. CLEMENT, New York, N. Y. Filed Dec. 8, 1921, issued Apr. 12, 1927. Assigned to Western Electric Co.

1,624,537—OSCILLATION GENERATOR—E. H. COLPITTS, East Orange, N. J. Filed Feb. 1, 1918, issued Apr. 12, 1927. Assigned to Western Electric Co.

1,624,562—ELECTRON DISCHARGE DEVICE—V. L. RONCI, Brooklyn, N. Y. Filed Dec. 18, 1923, issued Apr. 12, 1927. Assigned to Western Electric Co.

1,624,672—COMMUNICATION SYSTEM—H. W. O'NEILL, Elmhurst, N. Y. Filed Dec. 17, 1924, issued Apr. 12, 1927. Assigned to Western Electric Co.

1,624,673—METHOD OF AND MEANS FOR RADIO COMMUNICATION—A. E. PAIGE, Philadelphia, Pa. Filed Apr. 19, 1922, issued Apr. 12, 1927.

1,624,991 — RADIO TELEGRAPH AND TELEPHONE SYSTEM — C. SPEAKER, of Cherrydale, Virginia. Filed Jan. 29, 1921, issued Apr. 19, 1927.

1,624,966—AMBULATORY REPEATING SYSTEM—ROBERT W. MORRIS, of Roslyn, New York. Filed Aug. 25, 1923, issued Apr. 19, 1927. Assigned to American Telephone & Telegraph Co.

1,625,296—METHOD OF AND MEANS FOR MODULATING SIGNALING CURRENTS—EMORY LEON CHAFFEE, of Belmont, Mass. Filed Aug. 2, 1922, issued Apr. 19, 1927.

1,625,445—STATIC ELIMINATOR—J. ASTROM, et al, of Fort Wayne, Indiana. Filed Mar. 16, 1925, issued Apr. 19, 1927.

1,625,504—CONVERTER OF ELECTRIC CURRENT—F. G. SIMPSON, of Seattle, Washington. Filed July 6, 1926, issued Apr. 19, 1927.

1,625,776—ELECTRON-EMITTING CATHODE AND PROCESS OF PREPARING THE SAME—FREDERICK HOLBORN, of Hoboken, New Jersey. Filed Jan. 14, 1926, issued Apr. 19, 1927. Assigned to Hazeltine Corporation.

1,626,464—PORTABLE RADIO APPARATUS—W. M. HEINA, Bronx, New York. Filed Sept. 16, 1926, issued Apr. 26, 1927. Assigned to Heina Radio Corp.

1,626,526—CONDENSER—G. A. GILLEN, Jersey City, N. J. Filed May 7, 1924, issued Apr. 26, 1927. Assigned to Gillen, Kimmey Baker Syndicate, Inc.

1,626,537—VARIABLE CONDENSER—B. JIROTKA, Berlin, Germany. Filed Apr. 15, 1926, issued Apr. 26, 1927. Assigned to the Firm, Dr. Otto Sprenger.

1,625,822—RADIO RECEIVER AND TROUBLE ALARM—R. K. POTTER, of New York. Filed July 15, 1926, issued April 26, 1927. Assigned to American Telephone & Telegraph Co.

1,625,823—RADIO RECEIVER AND TROUBLE ALARM—R. K. POTTER, of New York. Filed July 15, 1926, issued Apr. 26, 1927. Assigned to American Telephone & Telegraph Co.

1,625,996—ELECTRICAL APPARATUS—L. O. GRONDAHL, of Pittsburgh, and Harry M. Ryder, of Forest Hills Borough, Pennsylvania. Filed Nov. 3, 1923. Assigned to The Union Switch & Signal Company of Swissvale, Pa.

1,626,030—SPARK GAP DEVICE—J. H. EASTMAN, Detroit, Michigan. Filed May 17, 1926, issued Apr. 26, 1927.

1,626,239—RECEIVING CIRCUIT—L. D. KELLOGG, Deerfield, Illinois. Filed Oct. 6, 1922, issued Apr. 26, 1927. Assigned to Kellogg Switchboard and Supply Co.

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1,626,356—FIXED CONDENSERS—L. T. RHOADES, Mont Clare, Pa. Filed Apr. 20, 1925, issued Apr. 26, 1927. Assigned to New York Coil Co.

1,626,370—RADIO RECEIVER—H. F. WORNER, Philadelphia, Pa. Filed Aug. 21, 1924, issued Apr. 26, 1927.

1,626,391—VARIABLE AIR CONDENSER—A. D. CARDWELL, Rockville Center, N. Y. Filed Sept. 26, 1922, issued Apr. 26, 1927. Assigned to The Allen D. Cardwell Mfg. Corp.

1,626,449—AUTOMATIC ANTENNA REGULATION—W. W. BROWN, Schenectady, N. Y. Filed Dec. 29, 1925, issued Apr. 26, 1927. Assigned to General Electric Co.

1,626,634—ELECTROSTATIC RADIOTUNER—J. SLEPIAN, of Swissvale, Pa. Filed Aug. 3, 1921, issued May 3, 1927. Assigned to Westinghouse Electric & Manufacturing Co.

1,626,684—CONDENSER—R. E. MARBURY, of Wilkinsburg, Pa. Filed Oct. 1, 1923, issued May 3, 1927. Assigned to Westinghouse Electric & Manufacturing Co.

1,626,685—ACTIVATION BY MEANS OF HYDROGEN FREE, CARBON BEARING GAS—J. W. MARDEN, of East Orange, THOMAS P. THOMAS and J. E. CONLEY, of Bloomfield, N. J. Filed Apr. 10, 1923, issued May 3, 1927. Assigned to Westinghouse Electric & Manufacturing Co.

1,626,699—RADIO TRANSMISSION SYSTEM—H. SCHIEFERSTEIN, of Charlottenburg, Germany. Filed Oct. 31, 1921, issued May 3, 1927. Assigned to Westinghouse Electric & Manufacturing Co.

1,626,724—FREQUENCY CONTROLLING SYSTEM—C. S. DEMAREST, of Ridgewood, N. J., and M. L. ALMQUIST, of Brooklyn, N. Y. Filed Dec. 31, 1923, issued May 3, 1927. Assigned to American Telephone & Telegraph Co.

1,626,803—SELF CONTAINED RADIO RECEIVING UNIT—J. S. FISHBACK, of Indianapolis, Ind. Filed Mar. 27, 1922, issued May 3, 1927.

1,626,858—VACUUM TUBE—H. E. METCALF, of San Leandro, Calif. Filed Feb. 28, 1924, issued May 3, 1927. Assigned to The Magnavox Co.

1,627,127—ELECTRIC DISCHARGE DEVICE—J. H. THOMPSON, of Hollidays Cove, W. Va. Filed Mar. 18, 1925, issued May 3, 1927.

1,627,231—DETECTOR AND METHOD OF CONTROLLING THE SAME—E. L. CHAFFEE, of Belmont, Mass. Filed Mar. 31, 1915, issued May 3, 1927. Assigned to John Hays Hammond, Jr.

1,627,411—ELECTRON RELAY—S. RUBEN, of New York, N. Y. Filed Oct. 23, 1920, issued May 3, 1927.

1,627,493—ELECTRICAL CONDENSER—W. DUBILIER, of New Rochelle, N. Y. Filed Feb. 28, 1925, issued May 3, 1927. Assigned to Dubilier Condenser Corporation.

1,627,707—RADIO SIGNALING SYSTEM—MARIUS LATOUR, of Paris, France. Filed Aug. 19, 1921, issued May 10, 1927. Assigned to Latour Corporation.

1,627,592—RADIO APPARATUS AND CURRENT CONTROL DEVICE THEREFOR—P. A. CHAMBERLAIN, of Chicago, Ill. Filed Dec. 4, 1926, issued May 10, 1927. Assigned to Mohawk Corp.

1,627,718—RADIO TELEPHONY—P. WARE, of New York, N. Y. Filed Sept. 3, 1921, issued May 10, 1927. Assigned to Ware Radio, Inc.

Re. 16,606—RADIO RECEIVING APPARATUS—H. F. LOWENSTEIN and E. E. CLEMENT, Washington, D. C. Original filed Aug. 14, 1922; Reissue Jan. 6, 1927, issued Apr. 26, 1927. Assigned to Edward F. Colladay.

1,625,234—RADIO APPARATUS—G. F. WITTGENSTEIN, Zurich, Switzerland. Filed Feb. 26, 1926, issued Apr. 19, 1927.

1,625,330—RADIO CONDENSER—A. S. PINKUS, New York City. Filed Apr. 10, 1925, issued April 19, 1927.

1,625,409—LIGHT SENSITIVE CELL CONTROL CIRCUIT—T. W. CASE, Auburn, N. Y. Filed Jan. 11, 1923, issued Apr. 19, 1927. Assigned to Case Research Laboratory, Inc.

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Maine,	Belfast, Box 296	Leighley, H. M.
Maryland,	Cumberland, 504 Fifth St.	Sweitzer, Guy H.
Massachusetts,	Arlington, 12 Harvard St.	Pristas, John E.
	Beverly, 132 Dodge St.	Colby, Edw. B.
	Medford, 51 Central Ave.	Wyman, R. C.
Michigan,	Detroit, 5526 3rd St.	Alguire, L. W.
	Detroit, 5711 Wabash Ave.	Cash, Geo. B.
	Detroit, Station WAFD, Addison Hotel	Lathrop, Fred A.
Missouri,	St. Joseph, 1810 Crescent Drive,	Abercrombie, J. B.
	St. Louis, 203 Pine St.,	Croxton, G. B.
Nebraska,	Sidney, 307 Central Ave.	Houge, Chas. A.
New Hampshire,	Manchester, 47 Amherst St.	Ryan, M. John
New Jersey	Camden, Victor Talking Mach. Co.	Gray, J. Calvin
	Elizabeth, Box 81 RFD No. 2	Hohner, John
	Fanwood, North Ave.	Hall, C. A.
	Fanwood,	Todd, Harold C.
	Ridgewood, 51 Highwood Ave.	Boyce, W. R. Jr.
New York,	Albany, 140 Grove Ave.	Thomas, J. D.
	Arvene (L. I.), 6603 Blvd.	Gradinger, Archie
	Brooklyn, 1831 East 34th St.	Gruetzke, C. P. E.
	Brooklyn, 21 Vermont Ave.	Kent, C. W., Jr.
	Brooklyn, 2021 East 13th St.	McLaughlin, C. E.
	Brooklyn, 3523 Bedford Ave.	Siegel, Ralph
	Brooklyn, 5007 Sixth Ave.	Slepian, Edward
	Buffalo, 110 Waverly St.	Clark, Chas. W.
	Buffalo, 98 Cayuga St.	Hedtke, Edward
	Hamilton, Box 31, Colgate University	Dowd, Alfred
	Lynbrook (L. I.) 4 Atlantic Ave.	Seidler, A.
	Mount Vernon, 14 North 9th Ave.	Dickely, Fred C.
	New York, 461 East 8th Ave.	Campbell, J. A.
	New York, Radio Inspector, Sub-Treasury Bldg.	Cochran, E. C.
	New York, care Postmaster, USS Worden	Colvin, Ruel
	New York, 16 Arden St.	Fernandez, Edw.
	New York, 124 West 42nd St.	Heuschkel, Paul J.
	New York, 785 Home St.	Jablon, W. W.
	New York, 113 West 57th St.	LeRoy, Frank S.
	New York, 224 West 13th St.	Taylor, Jay L.
	Ozone Park (L. I.), 8605 95th Ave.	Weilminster, Chas.
	Richmond Hill (L. I.), 9419 127th St.	Dowd, W. A.
	Rochester, 238 S. Goodman St.	Erhart, V. J.
	Rochester, 762 Bay St.	Friedler, D. P.
Ohio,	Cleveland, 2108 East 96th St.	Sherman, K. S.
	Cleveland, Radio Station WHK	Smith, S. E.
Oregon,	Portland, 109 East 42nd	La Salle-Dilley, L.
	Reedsport, Box S. P.	White, F. I.
Pennsylvania,	Oil City, 122 Washington Ave.	Hahn, S. L.
	Philadelphia, 1533 Pine St.	Keyser, Chas. S.
	Philadelphia, 1533 Pine St.	McClair, R.
	Philadelphia, 4223 Osage Ave.	Waltz, W. W.
	Philadelphia, 5010 Catherine St.	Yost, John C.
Texas,	Ft. Worth, 1601 Enderly Place	Abey, Robt. E.
	San Antonio, 1025 South Presa St.	Wall, L. D.
Washington,	Long View, P. O. Box 5, 466 19th Ave.	Trotter, C. A.
	Seattle, 116 Fairview Ave.	Ackerley, James
	Seattle, 703 25th Ave., N. W.	Blom, R. J.
	Seattle, 7708 Latona Ave.	Brandt, O. T. D.
	Seattle, 2049 E. Newton St.	Burleigh, John A.
	Seattle, 903 Telephone Bldg.	Eaden, Neal
	Seattle, 818 East 62nd St.	Flagler, R. H.
	Seattle, 903 Telephone Bldg.	Smith, Oliver C.
West Virginia,	Chester, 325 3rd St.	Faudree, R. F.
Wisconsin,	Milwaukee, 1460 Eighth Ave.	Kazmierowski, J. J.
Canal Zone,	Balboa, Box 35	Young, J. W.
Canada,	Nova Scotia, Halifax, 158 Granville St.	Baldwin, G. H.
	Nova Scotia, Halifax, USS Tampa	O'Day, Arthur C.
	Ontario, Westhill, Station CJYC	Carment, J. M
China,	HMS Hermes, China Station	Marshall, J. A.
Cuba,	Tuinucu,	Jones, Frank H.
England	Lincs., Skegness, Beam Radio Station	Towers, Robt.
Ireland,	Waterford City, No. 21 Shortcourse	Dunphy, Thos. O.
Mexico,	Mexico City, Jor Juana 119,	Veramendi, M.
	Tampico, Apartado 150	Cordova, F. B.

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Porto Rico,	Ensenada,	Schell, R. Jr.
S. W. Africa,	Walvis Bay, Radio Station	Fitzgibbon, M. B.
Spain,	Madrid, Mejia Lequerica 4	Moya, Miguel

Juniors Elected

California	Fresno, 1495 N. Van Ness Ave.	McKay, W. K.
	San Francisco, 790 Oak St.	Sphar, C. H.
Illinois,	Chicago, 7327 Phillips Ave.	Armstrong, H. W.
	Chicago, 1631 West 14th St.	Cohen, Theo.
	Deerfield, Radio Station WHT	Birkenhead, W. D.
	Rock Island, 1615 29½ St.	Foley, Jack
Massachusetts,	Boston, Jamaica Plains, 24 Beaufort Rd.	Goodrich, R. R.
	Winthrop, 88 Circuit Rd.	Colton, H. C.
New York,	Huntington Station (L. I.)	Flathman, Edw.
	New York, 100 318 W. 57th St.	Backlund, L. R.
	New York, 1075 Home St.	Rosenberg, I.
Pennsylvania,	Philadelphia, 5135 N. Fairhill St.	Mevius, John C.
	Zelienople, 302 E. New Castle St.	Bassler, S. G.
Texas,	Laredo, 216 Matamoros St.	Otal, Carlos, Jr.
Washington,	Everett, 2119 McDougal Ave.	Sands, L. G.
Canada,	Quebec, Montreal, 6 Delormier Ave.	Forward, Willis
	Quebec, Montreal, Bank of Montreal	Irving, E.
	Ontario, Sarnia, 430 S. Brock St.	Hunter, Alex.
	Ontario, Beamsville, RR No. 3	Potter, Robt. C.

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1,627,758—RADIO RECEIVING APPARATUS—MARCEL WALLACE, of New York, N. Y. Filed April 27, 1925, issued May 10, 1927. Assigned to Electro-Laboratories, Inc.

1,627,767—TRANSMISSION OF ELECTRIC OSCILLATIONS—L. N. BRILLOUIN & E. M. F. FROMY, of Paris, France. Filed Aug. 10, 1925, issued May 10, 1927.

1,627,815—RADIO BATTERY CHARGER—C. T. WEIBLER, of La Grange, Ill. Filed Nov. 24, 1926, issued May 10, 1927. Assigned to All-American Radio Corp.

1,628,009—SIGNAL TRANSMISSION SYSTEM—A. H. TAYLOR, of Washington, D. C. Filed Sept. 20, 1926, issued May 10, 1927. Assigned to Wired Radio, Inc.

1,628,045—ELECTRONIC DISCHARGE DEVICE—W. F. HENDRY, of Ossining, N. Y. Filed Oct. 8, 1926, issued May 10, 1927. Assigned to Manhattan Electrical Supply Co., Inc.

1,628,115—ELECTRIC CIRCUIT CLOSER FOR RELAYS—L. L. CALL, of Detroit, Michigan. Filed Apr. 17, 1922, issued May 10, 1927.

1,628,411—SYSTEM OF SECRET WIRELESS TELEPHONY—F. W. KRANZ, of Geneva, Ill. Filed Feb. 1, 1923, issued May 10, 1927.

1,628,466—ELECTRON DEVICE AND METHOD OF OPERATING—A. W. HULL, of Schenectady, N. Y. Filed Nov. 15, 1921, issued May 10, 1927. Assigned to General Electric Co.

1,628,627—ELECTRICAL STRUCTURE—WM. DUBILIER, of New York, N. Y. Filed June 30, 1923, issued May 10, 1927. Assigned to Dubilier Condenser Corp.

1,628,648—RADIO TELEGRAPHIC STATION—J. BETHENOD, of Paris, France. Filed Aug. 2, 1921, issued May 17, 1927.

1,628,659—ELECTRICAL CONDENSER—C. S. FRANKLIN, of London, England. Filed June 21, 1921, issued May 17, 1927. Assigned to Radio Corp. of America.

1,628,676—RECEIVING ARRANGEMENT FOR WIRELESS TELEGRAPHY—A. MEISSNER, of Berlin, Germany. Filed Sept. 3, 1921, issued May 17, 1927. Assigned to Gesellschaft Fur Drahtlose Telegraphie M. B. H.

1,628,982—ELECTRON DISCHARGE DEVICE—R. I. HULSIZER, of East Orange, N. J. Filed May 28, 1921, issued May 17, 1927. Assigned to Western Electric Co.

1,628,905—WIRELESS TELEPHONE SYSTEM—A. NYMAN, of Wilkinsburg, Pa. Filed Sept. 2, 1921, issued May 17, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,628,987—THERMIONIC DEVICE AND METHOD OF EVACUATING THE SAME—R. W. KING, of New York, N. Y. Filed Sept. 24, 1920, issued May 17, 1927. Assigned to Western Electric Co.

1,628,999—ELECTRON DISCHARGE DEVICE—V. L. RONCI, of Brooklyn, N. Y. Filed Dec. 19, 1924, issued May 17, 1927. Assigned to Western Electric Co., Inc.

1,629,001—OSCILLATION GENERATOR—J. C. SCHELLENG, of East Orange, N. J. Filed Sept. 18, 1923, issued May 17, 1927. Assigned to Western Electric Co., Inc.

1,629,009—LOW IMPEDANCE ELECTRIC DISCHARGE DEVICE—H. C. SNOOK, of South Orange, N. J. Filed Aug. 7, 1920, issued May 17, 1927. Assigned to Western Electric Co. Inc.

1,629,020—CONDENSER—E. B. CRAFT, of Hackensack, N. J. Filed Sept. 6, 1922, issued May 17, 1927. Assigned to Western Electric Co.

1,629,337—RADIO FIRE ALARM SYSTEM—H. GARRETT, of Dallas, Texas. Filed Oct. 14, 1925, issued May 17, 1927.

1,629,171—ELECTRON DISCHARGE DEVICE—A. MAVROGENIS, of Milwaukee, Wis. Filed Sept. 16, 1926, issued May 17, 1927.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED JUNE 1, 1927

Transferred to Fellow Grade

Dist. of Columbia,	Washington	Hund, August

Transferred to Member Grade

Massachusetts,	E. Springfield	Laport, E. A.
Canada,	Montreal, c/o Radio La Presse,	Spencer, L.

Elected to Member Grade

Canada,	Toronto, Hydro Elec. Power Comm.	Dalton, F. K.
England,	Liverpool, 76 Old Hall Street,	Lyons, C. L.

Associates Elected

Alabama,	Pratt City, 58 Vine St.	Hassler, W. D.
Arizona,	Jerome, Box 726	Rutherford, R. L.
California,	Fresno, P. O. Box 726	Wells, Frank T.
	Hollywood, 6533 Hollywood Blvd.	Drury, Paul O.
	Long Beach, 216 Grand, Apt. A.	Lanphere, Merle
	Los Angeles, 1707 So. Flower	Baker, Lewis C.
	Los Angeles, 2718 Moss Ave.	Kemper, Guy A.
	Los Angeles, 5120 So. Wilton Place	Rusthol, H. W.
	Los Angeles, 331 S. Mariposa Ave.	Vauchelet, B. F.
	Los Angeles, 5753 8th Ave.	Walling, H. S.
	Los Angeles, 1020 Santa Fe Ave.	Woods, Edw. J.
	Oakland, 5835 College Ave.	Burchard, Bruce M.
	Riverside, 309 Lime St.	Chapman, C. W., Jr.
	Riverside, 176 Riverside Drive	West, Lloyd E.
	San Francisco, 140 New Montgomery St.	Parker, Ray H.
	San Francisco, 1869 Stockton St.	Stagnaro, John A.
	Whittier, 410 N. Milton	Brown, Wilbur C.
	Whittier, Box 103,	Mendeck, Fred C.
Connecticut,	Torrington, 150 Oak Ave.	York, P. J.
Florida,	Miami, 56 S. E. First St.	Durante, J. A.
	St. Augustine, 10 Sevilla St.	McCallum, F. L.
	Tampa, 339 Plant Ave.	Moore, Wm. P.
Georgia,	Atlanta, 102 West Peachtree St.	Alexander, D. C.
	Atlanta, 963 Ponce de Leon	Anderson, L. E.
	Atlanta, 533 Boulevard Place, N. E.	Bayliss, V. B.
	Atlanta, 413 6th St.. N. E.	Bristol, T. R.
	Atlanta, 690 Piedmont Ave., N. E.	Cook, J. M.
	Atlanta, 902 Georgia Savings Bank Bldg.	Cota, John J.
	Atlanta, Station WSB, Atlanta Journal	Daugherty, C. F.
	Atlanta, 1010 E. Shadow Lawn Ave.,	Dendy, G. R.
	Atlanta, 69 Mangum St., S. W.	Foote, James J.
	Atlanta, 123 Spring St.,	Hill, Arthur M.
	Atlanta, 843 N. Boulevard,	Kay, Lambdin
	Atlanta, A. T. & T., Hurt Bldg.,	Keith, John M.
	Atlanta, 524 Post Office Bldg.	Llewellyn, Geo.
	Atlanta, 16 Auburn Ave.	Lyon, D. P.
	Atlanta, 8 Peachtree Way	MacMillan, R. H.
	Atlanta, 1765 N. Decatur Road	Morris, Jas. S.
	Atlanta, 12 Walton St.	Mosteller, M. O.
	Atlanta, 902 Flatiron Bldg.	Murphy, Brower
	Chambler, 8th & Jackson Sts.	Mindenhall, W. W.
	East Point, 423 E. Washington Ave.	Jones, J. C.
Illinois,	Chicago, 1662 Clifton Park Ave.	Goldman, Stanley
	Chicago, 7526 Robey St.	Mooney, Raymond
	Oak Park, 1526 N. Austin Blvd.	Witt, Harold A.
Iowa,	Dubuque, care General Delivery	Roedell, L. W. F.
	Muscatine, 919 Iowa Ave.	Wildman, Thos. S.
Kansas,	Great Bend, 1612 Murphy St.	Mitchell, Theo. R.
Kentucky,	Bypro,	McGeorge, D. H.
	Louisville, P. O. Box 855	Berry, Robt. C.
Louisiana,	New Orleans, 841 Carondolet St.	Andres, Chas. Jr.

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[illegible]	Belfast, Box 296	Leighley, H. M.
[illegible]	Cumberland, 504 Fifth St.	Sweitzer, Guy H.
Massachusetts,	Arlington, 12 Harvard St.	Pristas, John E.
	Beverly, 132 Dodge St.	Colby, Edw. B.
	Medford, 51 Central Ave.	Wyman, R. C.
Michigan,	Detroit, 5526 3rd St.	Alguire, L. W.
	Detroit, 5711 Wabash Ave.	Cash, Geo. B.
	Detroit, Station WAFD, Addison Hotel	Lathrop, Fred A.
Missouri,	St. Joseph, 1810 Crescent Drive,	Abercrombie, J. B.
	St. Louis, 203 Pine St.,	Croxton, G. B.
Nebraska,	Sidney, 307 Central Ave.	Houge, Chas. A.
New Hampshire,	Manchester, 47 Amherst St.	Ryan, M. John
New Jersey	Camden, Victor Talking Mach. Co.	Gray, J. Calvin
	Elizabeth, Box 81 RFD No. 2	Hohner, John
	Fanwood, North Ave.	Hall, C. A.
	Fanwood,	Todd, Harold C.
	Ridgewood, 51 Highwood Ave.	Boyce, W. R. Jr.
New York,	Albany, 140 Grove Ave.	Thomas, J. D.
	Arverne (L. I.), 6603 Blvd.	Gradinger, Archie
	Brooklyn, 1831 East 34th St.	Gruetzke, C. P. E.
	Brooklyn, 21 Vermont Ave.	Kent, C. W., Jr.
	Brooklyn, 2021 East 13th St.	McLaughlin, C. E.
	Brooklyn, 3523 Bedford Ave.	Siegel, Ralph
	Brooklyn, 5007 Sixth Ave.	Slepian, Edward
	Buffalo, 110 Waverly St.	Clark, Chas. W.
	Buffalo, 98 Cayuga St.	Hedtke, Edward
	Hamilton, Box 31, Colgate University	Dowd, Alfred
	Lynbrook (L. I.) 4 Atlantic Ave.	Seidler, A.
	Mount Vernon, 14 North 9th Ave.	Dickely, Fred C.
	New York, 461 East 8th Ave.	Campbell, J. A.
	New York, Radio Inspector, Sub-Treasury Bldg.	Cochran, E. C.
	New York, care Postmaster, USS Worden	Colvin, Ruel
	New York, 16 Arden St.	Fernandez, Edw.
	New York, 124 West 42nd St.	Heuschkel, Paul J.
	New York, 785 Home St.	Jablon, W. W.
	New York, 113 West 57th St.	LeRoy, Frank S.
	New York, 224 West 13th St.	Taylor, Jay L.
	Ozone Park (L. I.), 8605 95th Ave.	Weilminster, Chas.
	Richmond Hill (L. I.), 9119 127th St.	Dowd, W. A.
	Rochester, 238 S. Goodman St.	Erhart, V. J.
	Rochester, 762 Bay St.	Friedler, D. P.
Ohio,	Cleveland, 2108 East 96th St.	Sherman, K. S.
	Cleveland, Radio Station WHK	Smith, S. E.
Oregon,	Portland, 109 East 42nd	La Salle-Dilley, L.
	Reedsport, Box S. P.	White, F. I.
Pennsylvania,	Oil City, 122 Washington Ave.	Hahn, S. L.
	Philadelphia, 1533 Pine St.	Keyser, Chas. S.
	Philadelphia, 1533 Pine St.	McClair, R.
	Philadelphia, 4223 Osage Ave.	Waltz, W. W.
	Philadelphia, 5010 Catherine St.	Yost, John C.
Texas,	Ft. Worth, 1601 Enderly Place	Abey, Robt. E.
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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 15 AUGUST, 1927 Number 8

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GENERAL INFORMATION

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INSTITUTE ACTIVITIES

Personal Mention

Dr. J. H. Dellinger, Junior Past President of the Institute, recently sailed for Europe where he will be traveling for three months in the interest of the Department of Commerce, particularly to study radio aids to navigation in other countries.

Mr. L. E. Whittemore, member of the Board of Direction of the Institute, has been appointed Secretary of the American Delegation to the International Radio Telegraph Conference to be held in Washington, D. C., in October. Mr. Whittemore is now in Washington where he will remain until the conclusion of the Conference. His address is American Delegation, International Radio Telegraph Conference, United Chamber of Commerce Building, 1615 H Street, N. W., Washington.

During Mr. Whittemore's absence from New York, Professor L. A. Hazeltine is serving as Acting Chairman of the Institute's Committee on Standardization.

News of the Sections

Atlanta Section

The first formal meeting of the newly organized Atlanta Section was held in the Atlanta Athletic Club on the evening of July 6th. Major Walter Van Nostrand presided.

The major portion of the meeting was devoted to permanent organization work. The officers of the Section were elected as follows: Chairman, Major W. Van Nostrand; Vice-Chairman, H. L. Wills; Secretary-Treasurer, George Llewellyn; and Assistant Secretary-Treasurer, John M. Keith. The Chairmen of the Committees are: Meetings and Papers, H. L. Wills; Publicity, Lambdin Kay; Membership, C. F. Daugherty.

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Following the business session, the members present informally discussed the effect on propagation of waves under varying power using Hertzian antennas.

At a meeting of the Atlanta section held on August 3rd, Major W. Van Nostrand presiding, Ray F. Lovelee, of the Federal Radio Corporation, presented a paper entitled "Broadcast Receiver and Power Supply Designed."

A general discussion followed the presentation of this paper.

There were twenty members of the Institute present at the meeting and at the business meeting and dinner preceding.

The next meeting of the Atlanta section will be held on September 7th in the Atlanta Athletic Club.

Philadelphia Section

A meeting of the Philadelphia Section was held on June 24, 1927 in the Franklin Institute. J. C. Van Horn was the presiding officer.

A paper was presented by C. Francis Jenkins on "Radio Vision." Prof. Lloyd M. Knoll, John Arnold and others participated in the discussion which followed.

Due to his removal to New York City, the resignation of David P. Gullette as Secretary of the Section was accepted. John C. Mevius was elected to succeed Mr. Gullette.

The attendance at this meeting was over one hundred.

The next meeting of the Philadelphia Section will be held on September 23, 1927 at which time Prof. Charles M. Weyl will present a paper on "Reproduction of Sound."

Rochester Section

On May 20, 1927 at a meeting of the Rochester Section held in the Sagamore Hotel, Benjamin Olney of the Stromberg Carlson Telephone Manufacturing Company delivered a paper entitled "Acoustics as Related to Radio Engineering."

Following the technical portion of the meeting, the following officers were elected for the next Section-year by unanimous vote: Chairman, H. J. Klumb; Vice-Chairman, A. E. Soderhon; Secretary-Treasurer, F. Reynolds. Mr. J. F. Hitchcock was elected Chairman of the Board of Direc-

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tion of the Section. The following are Chairmen of the standing Committees: Meetings and Papers, V. M. Graham; Constitution, A. E. Roeser; Membership, A. L. Schoen; Rooms, A. E. Soderhon.

Seattle Section

A meeting of the Seattle Section was held on May 11, 1927 in the Club Room of the Telephone Building, Seattle.

T. M. Libby was the presiding officer. Two papers were read. The first, by Charles L. Reynolds, was on "Our Patent Laws," covering in detail the history and development of our present laws together with an account of some of the requirements for valid patents.

The second paper was presented by E. S. Smith on "Cathode Ray Oscillographs." This paper described the development, construction and use of both hot and cold cathode ray oscillgraphs. The paper was illustrated with slides of equipment, setups and the resulting oscillograms.

At a meeting of the Seattle Section on June 11, 1927, two papers were presented. Charles Peterson delivered a paper on "Loud Speaking Receiver Design," describing the basic design and manufacturing problems of loud speaking receivers.

The second paper, by J. A. Burleigh and O. C. Smith was on the subject of "Sounds We Can and Cannot Hear," which included a striking demonstration of the frequency-response characteristics of the human ear.

The next meeting of the Seattle Section will be held early in September.

I. R. E. Notes

Nominating Committee

For the purpose of recommending to the Board of Direction a program of possible candidates for the offices of President, Vice-President and two Directors of the Institute a nominating committee has been appointed by President Bown. The membership of this committee is as follows: J. V. L. Hogan, Chairman; L. A. Hazeltine and Melville Eastham. The Committee desires to receive suggestions from the Institute membership.

As authorized in the amendment to the Institute's Constitution dated January 1921, "Nomination by Petition shall

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be made by letter, addressed to the Board of Direction, setting forth the name of the proposed candidate and the office for which it is desired he be nominated. For acceptance, a letter of Petition must reach the Board of Direction on or before October 15th of any year, and shall be signed by at least thirty-five Fellows, Members or Associates."

Changes of Address

It is very important that the office of the Institute be notified promptly of any change in address of any members of the Institute. Such notification should contain the following information: the former address, the new mailing address and the new business address (if the mailing address is a residential one.)

The Institute cannot be held responsible for misdirected mail if it is not notified at least two weeks prior to the publication date of the PROCEEDINGS.

Member and Fellow Diplomas

Members and Fellows of the Institute desiring a diploma of membership are requested to advise the office of the Institute the exact reading of the name they desire to have placed on the diploma.

Diplomas for Members or Fellows can be secured free of charge upon application to the Institute.

Conference of Engineering Societies Secretaries

At the third Conference of Secretaries of Engineering Societies held on June 16th and 17th in Cleveland, Ohio, R. A. Carlton, a member of the Cleveland Section of the Institute, represented the Institute. Much information of value was secured at this Conference.

Committee Work

I. R. E. Subcommittee on Power Supply

At a meeting of the Institute's Subcommittee on Power Supply of the Committee on Standardization, held at Institute Headquarters on July 13, 1927, the following committee members were present: W. E. Holland, Chairman; L. Baker, C. T. Burke, L. W. Chubb, R. A. Klock, E. J. Moore, G. E. Rodwin, G. W. Vinal.

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This was the first meeting of the Committee and, despite the summer heat, the Committee remained in session all day and made good progress on the following subjects:

a. Terms and definitions applying to radio socket-power units.

b. Standard method of ripple measurement.

c. Method of measuring inductance of filter choke coils.

d. Method of measuring capacitance and power factor of filter condensers.

Items a, b, and c have been discussed previously in meetings of the Socket Power Committee of the National Electric Manufacturers' Radio Division and have been referred recently to the Institute's Power Supply Subcommittee by the N. E. M. A. Committee as being work that should more properly be undertaken by the Institute.

The Radio Manufacturers' Association Engineering Division is also looking to the Institute's Power Supply Subcommittee to work out standards on the above subjects.

The Subcommittee plans to meet again in September. It is hoped that it will then be able to make final recommendations to the Committee on Standardization on all four of the above subjects.

Sectional Committee on Radio, A. E. S. C.

Committee on Vacuum Tubes

A meeting of this Committee was held at the office of the Institute of Radio Engineers. The meeting considered the standardization of tube bases, the preparation of information for the American representative on the Advisory Committee on Tube Base Standardization at the International Electrotechnical Commission meeting in Italy in September, and filament voltage tolerances.

Two types of UX base and tube terminology were adopted for recommendation.

The Committees on Transmitting and Receiving Sets and Installations, Component Parts and Wiring, Electro-Acoustic Devices and Power Supply and Outside Plant have reported no further activities during the month of May.

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OBITUARY

With deepest regret the Institute of Radio Engineers announces the death of

Cornelius Johannes de Groot

Dr. de Groot was born in den Helder, Holland in 1883, and was in succession a student at the Technical College, Delft, Holland and the Technical College at Karlsruhe, Germany. For many years he was a Fellow of the Institute of Radio Engineers.

He was one of the pioneers both in the long wave and short wave field of radio communication, was responsible for the installation of the highly modern radio stations in the Dutch East Indies (which communicate directly with Holland and other parts of the world), and has made many fundamental studies of atmospheric disturbances and transmission phenomena which he has published in brilliant papers in the "Proceedings of the Institute of Radio Engineers", and elsewhere. His careful observations and ingenious deductions and conclusions from them have been repeatedly verified by later investigations.

He will unquestionably be remembered as a pioneer worker and inspired investigator, one who brought the benefits of radio communication to the distant parts of the earth, and a most likeable gentleman.

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A NEW FREQUENCY TRANSFORMER OR FREQUENCY CHANGER

By
Isaac Koga

Many devices have been published to multiply the frequency of a given alternating current, but there is no static means to obtain 1/2, 1/3 etc. of the given frequency. This paper explains how such frequencies can be obtained by means of a three-electrode vacuum tube. The same scheme can also be used as a frequency multiplier, and even to get frequencies such as 3/2, 2/3, 4/3, 3/4 etc. of a given frequency.

The process is quite simple; to get for example half the frequency of a given alternating current of frequency f_o, construct any valve oscillation generator having natural frequency of about $\frac{1}{2} f_o$ and excite its grid or plate circuit with

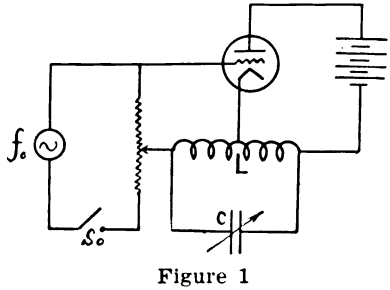

Figure 1

proper voltage of frequency f_o, the resulting frequency will be just $\frac{1}{2} f_o$. The process of getting any other frequencies having simple ratios to the given frequency is quite similar, so that I shall describe only a few cases in order to give a general idea of the method.

Take, for example, the Hartley circuit (Figure 1), the grid of which can be excited by an external source by means of a potentiometer or a transformer. Without external exci-

Received by the Editor June 8, 1927.

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tation the frequency f of the oscillatory current is given by the natural frequency f_n of the circuit.

$$f = f_n = \frac{1}{2\pi\sqrt{CL}}$$

Let now an exciting voltage of f_c ($\fallingdotseq 2 f_n$) be gradually increased. At first, at very low excitation, the frequency of the oscillatory current is independent upon the exciting frequency f_o and is maintained at the value f_n, but as the exciting voltage increases, the frequency of the system begins to approach $\frac{1}{2} f_o$ until it suddenly coincides with $\frac{1}{2} f_o$,

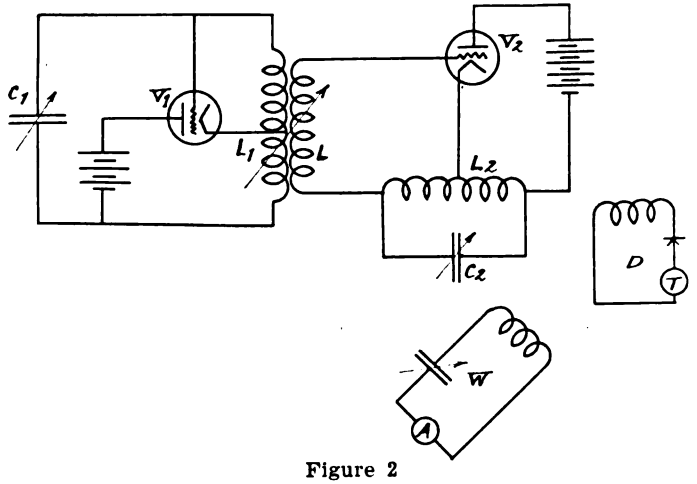

Figure 2

thereby causing sudden depression of plate current at the same time. The necessary exciting voltage may be lower, the smaller the difference of $\frac{1}{2} f_o$ and f_n. Next, keeping the exciting voltage constant, vary the natural frequency of the circuit in one direction, the oscillator frequency f falls suddenly in step with $\frac{1}{2} f_o$ and again suddenly out of step at a certain value of $\frac{1}{2} f_o \sim f_n$. This "attraction" of frequencies is already treated by E. V. Appeleton* in the case of two nearly equal frequencies, and the similar curves as in Figure 9 in his paper were also taken for our cases.

The oscillatory current thus produced in the oscillation circuit contains only a small amount of second harmonic of

*"Automatic Synchronization of Triode Oscillators", Cambridge Phil. Soc., Proc. 21 pp. 231-248, Nov., 1922.

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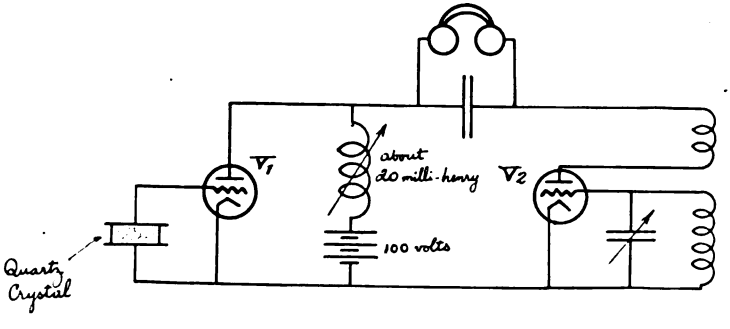
$\frac{1}{2} f_o$, i. e. the component of f_o itself, and the wave form seems to be sinusoidal on inspection.

If an excessive excitation, the whole system will be evidently forced to produce a current of frequency f_o; and if the difference $\frac{1}{2} f_o \sim f_n$ is very large, the system will take another frequency, automatically selecting the nearest to f_n which is yet in a simple ratio to f_o.

In Figure 2, V_1 is used as an exciter, V_2 as an oscillator which is controlled by V_1, both being Radiotrons UX-201-A. The frequency meter W is loosely coupled to the oscillation circuit $L_2 C_2$ only. Adjusting C_2 slowly to and fro, a beat note is heard in the crystal detector circuit D which starts and stops suddenly without giving a tone of low pitch. This



V1 V2 — Radiotron UX-201-A

Figure 3

shows that the oscillator V_2 suddenly falls into step and out of step with the frequency of V_1 in some simple numerical ratio. The frequency meter W gives two frequencies for which a beat is heard; one the frequency f_o of V_1 itself and the other a frequency f slightly different from the natural frequency f_n of V_2. (This change of frequency f was already noted). But when the beat note stops, only one frequency is observed and practically no trace of f_o in the corresponding setting of the frequency meter, which shows that there is now scarcely any component of f_o in the oscillator V_2. One of my experimental results is shown in the following table, the third column gives the natural frequency f_n of V_2 corresponding to the both limits just before the beat notes are heard under excitation. All numbers are given in kilocycles per second.

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Frequency of V_1	Frequency of V_2	Natural Frequency of V_2
544 x 2	544	542 - 551
538 x 3	538	535 - 544
436 x 4	436	432 - 438
445 x 5	445	443 - 448
373 x 6	373	372.5 - 374.5
321 x 7	321	320.5 - 321.5

As a practical application of this apparatus we tried to obtain a constant radio frequency oscillation of long wavelength using a quartz crystal controlled oscillator. In Figure 3, V_2 is a triode controlled by a crystal oscillator V_1. The fundamental frequency of the crystal oscillator is about 70

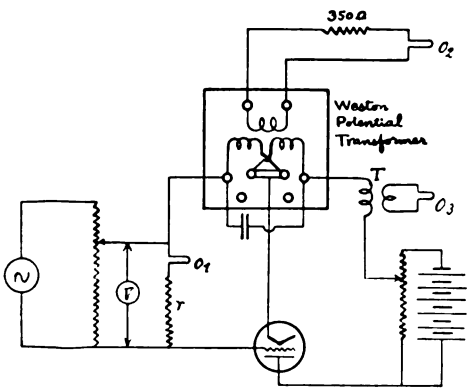

Figure 4

kilocycles per second and V_2 could be controlled easily down to one-tenth of this frequency i. e. about 7 kilocycles per second. If we use this process in a series of steps, we will be able to get still lower frequencies.

The wave form of the current in the various parts of the circuit and the manner of falling into step were observed by means of a Siemens electromagnetic oscillograph. Audio frequency alternating current was employed for this experiment, a potential transformer (Weston Potential Transformer Model 311, ratios 440 and 220 volts to 110 volts) was used as an inductance in the oscillatory circuit, the exciting voltage being secured from an inductor type alternating current generator designed for 500 cycles at full

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speed. The general connection arrangement is shown in Fig. 4. A resistance of about 350 ohms is inserted in the output circuit as a load. O_1 O_2 O_3 are oscillograph vibrators, T is a current transformer, and r indicates a series resistance. Figures 5-13 are picked out from about sixty oscillograms which were secured.

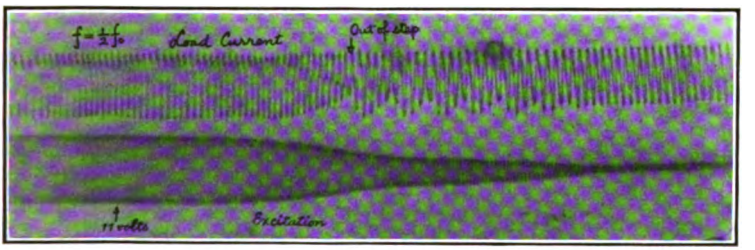

Figure 5

Figure 5 shows the oscillograms of the exciting voltage and load current. The excitation is decreased from 11 volts to zero. "Synchronization" (which is the term I shall employ for the sake of simplicity for the condition when the two oscillators fall into step with a definite frequency ratio) fails to occur below about 7 volts. With no excitation the frequency of the oscillator f is f_n

$$f = f_n \div \frac{2}{5} f_0 ,$$

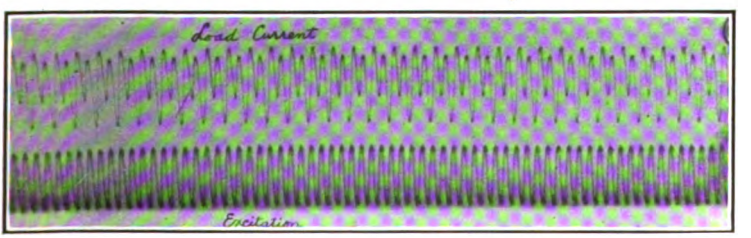

Figure 6

but over 7 volts excitation, we have

$$f = \frac{1}{2} f_0 ,$$

and over 11 volts the load current contains considerable amount of exciting frequency component; and still further

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increase in exciting voltage gives only f_o, and the component of frequency $\frac{1}{2} f_o$ disappears.

Figures 6 and 7 cover the case when the excitations are slightly above and below 11 volts respectively. Figure 8 shows the case when the possibility of free oscillation is altered by decreasing the plate voltage, the excitation being

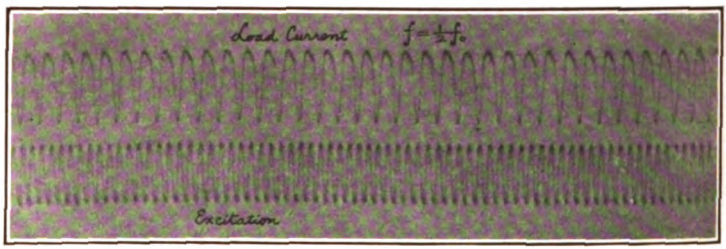

Figure 7

kept at about four volts. Alternating current of 50 cycles is also shown as a timing wave. With certain plate voltage, the oscillatory current cannot be maintained at a fractional value of the given frequency, and the system acts only as a poor amplifier. From all governing considerations, it is probably certain that this type of frequency control cannot

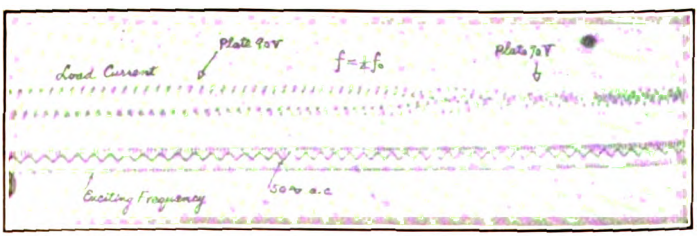

Figure 8

be carried out unless the oscillatory circuit satisfies the condition of self-oscillation.

Sometimes when

$$\frac{1}{2} f_o - f_n > f_n - \frac{1}{3} f_o ,$$

the circuit is sometimes controlled to a frequency of 1/3 f_o at

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certain exciting voltage, and then to $\frac{1}{2} f_0$ at a little higher voltage. Figure 9 is another example when the oscillator is forced into the relation

$$f = \frac{1}{4} f_0 .$$

The excitation is begun from 14 volts, and lowered to zero.

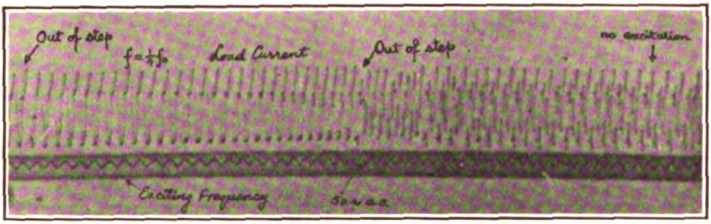

Figure 9

The load current is somewhat complicated but the general characteristics are nearly the same as those of Figure 5 as a whole. Figure 10 shows the plate current and load current when

$$f = \frac{1}{6} f_0 .$$

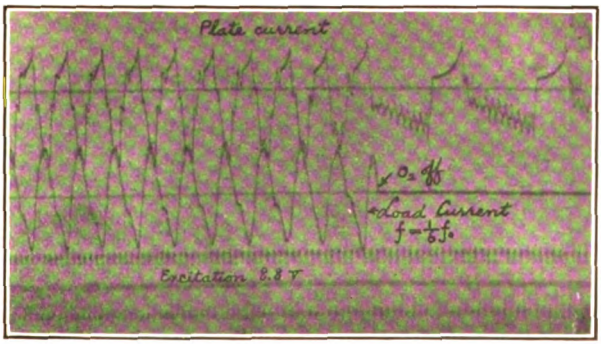

Figure 10

The load being cut off midway, the natural frequency of the oscillator decreases with the increase of inductance, the plate current is distinctly seen to contain the component of exciting frequency. Evidently the synchronization fails to occur under no-load because the oscillation becomes too

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strong for the same controlling voltage. Similar oscillograms were taken up to $f = \frac{1}{8} f_0$ which are omitted here.

On the other hand when the frequency of excitation f_0 is lower than the natural frequency f_n of the oscillatory circuit, the amplitude of load current is generally very small

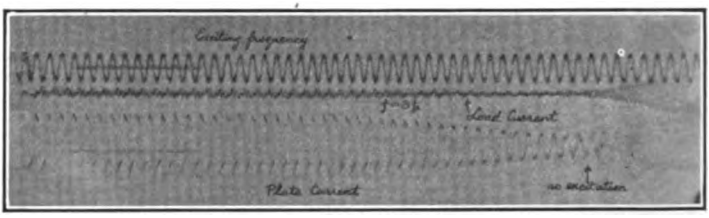

Figure 11

compared with that in the foregoing cases. That the plate current has the exciting frequency as its fundamental frequency is again quite a different circumstance from that in the foregoing cases. In Figure 11 the grid excitation is decreased from 15 volts to zero. It is seen that while the load current is depressed markedly when controlled, it becomes of considerable amplitude when the excitation is removed. The frequency relations are:

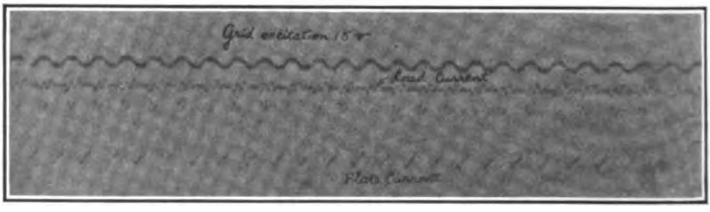

Figure 12

$$f = 3 f_0 \text{ while } f_n \div \frac{7}{2} f_0$$

Figures 12 and 13 show the cases when the circuit is controlled so that

$$f = 4 f_0 \text{ and } f = 9 f_0$$

These phenomena of synchronization can be explained as due to the "attraction" of two nearly equal frequencies, and to the non-linear characteristic of a vacuum tube. Sup-

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pose that the exciting frequency f_0 and the frequency f of the oscillator can be expressed as:

$$f_0 = \frac{n}{m} F, \quad f = F + \epsilon .$$

where F is a fixed value,

m and n are any integers, not including zero,

ϵ is a very small quantity, (positive or negative), compared to F.

As the grid voltage-plate current characteristic is not linear, if the exciting voltage of $\frac{n}{m} F$ is superposed, the current in the oscillatory circuit will be composed of several

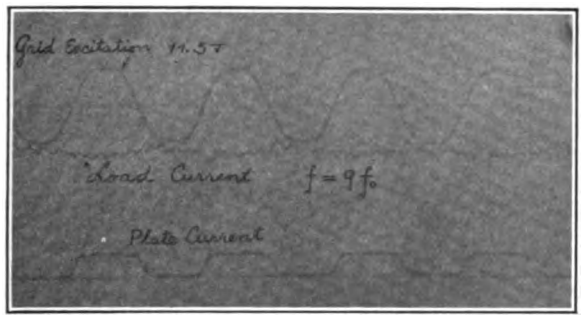

Figure 13

frequencies, among which the following three frequencies will be predominant:

$$F + (n+1)\epsilon, \quad F - (n-1)\epsilon, \quad F + \epsilon.$$

The first two frequencies are the beat frequencies between the m th harmonic of the exciting frequency and the $(n \pm 1)$ th harmonics of the oscillator frequency.

In practice n is ordinarily small, so that the most important components are the last two.

Now, as the exciting voltage increases, these three frequencies cannot coexist independently, and they "attract" each other. But as the value F is fixed by the external force, these three frequencies necessarily become F, thereby reducing the value of ϵ to zero.

It will be sometimes convenient to give a proper name to this new device. It is believed that the role of this apparatus is quite analogous to that of an ordinary step-down transformer. Just as the latter offers a means of obtaining

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any lower voltage (or current) proportional to the input voltage (or current), so the former provides a means of obtaining a lower frequency which holds any required ratio to the given frequency. From this viewpoint, I propose the name "(Static) Frequency Transformer", for the device. Considered, however, from another point of view, this apparatus gives a frequency having any given ratio to the original frequency, so that the name "(Static) Frequency Changer" is also proposed.

SUMMARY: Alternating current having a frequency which is any fractional value of the frequency of a supplied current, can be obtained by means of a triode oscillator. The phenomenon seems to be due to the "attraction" of two nearly equal frequencies occurring in the triode circuit and to the non-linear characteristic of a triode.

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AUDIO FREQUENCY TRANSFORMERS*

By
JOHN M. THOMSON

The following assumptions will be made with respect to the characteristics of the transformer.

The primary and secondary inductances are independent of frequency. This is true as long as the flux is uniformly distributed across the core. The inductance will vary with the amount of direct current in the primary, but for each value of direct current the inductance will be constant. Actually the inductance will vary with the strength of the signal. Therefore the amplification will depend on the signal. As the component of the self inductance due to the flux which passes through the coil itself is small the unequal distribution of current across the conductor will have very little effect on the inductance.

The resistances of the primary and secondary coils are independent of frequency. The correctness of this assumption will depend on the size of the conductors used and on the length of the leakage paths. Using the methods developed for power transformers the increase in the resistance of the transformers made by the company with which the writer is associated was found to be less than 1 per cent at 10,000 cycles.

The mutual capacity was taken equivalent to a lumped capacity connected across the terminals. Whether or not this is the correct position for the equivalent capacity is open to discussion.

The self capacity is also assumed to be equivalent to a lumped capacity connected across the terminals of the secondary. The mutual and self capacities are assumed to be independent of frequency but will most likely vary slightly with the frequency if the conductors and the length of the

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leakage paths are such as to give a non-uniform distribution of current.

The exciting current and the core loss have been neglected, but could be represented as in power transformer work by an impedance shunting the primary.

Transient phenomena are neglected.

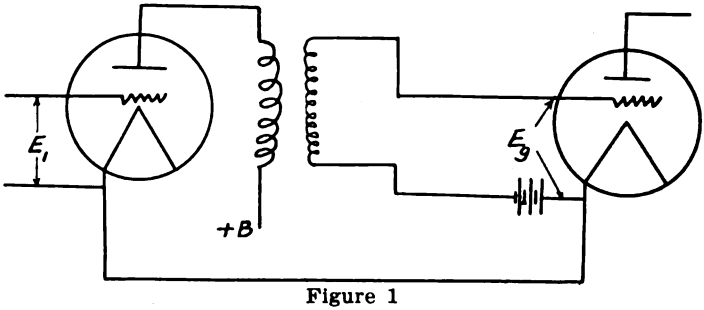

Figure 1

R_P = tube plate impedance in ohms.
U = tube amplification factor = μ, in Figures 5 and 6.
R_1 = resistance of the primary in ohms.
R_2 = resistance of the secondary in ohms.
L_1 = inductane of the primary in henrys.
L_2 = resistance of the secondary in henrys.
M = mutual inductance in henrys.
C_m = mutual capacity between primary and secondary coils in farads.
C_1 = by-pass condenser across the primary plus the self capacity of the primary in farads.

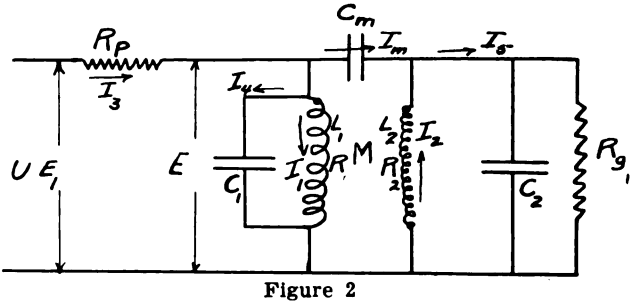

Figure 2

C_2 = Self capacity of the secondary in farads; plus the grid-filament capacity of the tube connected to the secondary of the coil.

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$w = 2\pi \times$ frequency in cycles per second.

$$Z_g = \frac{1}{\frac{1}{R_g} + \frac{1}{\frac{1}{j w c_2}}} = \frac{R_g}{1 + j w c_2 R_g}$$

$$Z_1 = R_1 + j w L_1$$

$$Z_2 = R_2 + j w L_2$$

The connection of the transformer in the set is given in Fig. 1. Fig. 2 gives the equivalent a-c. electrical circuit.

The following equations were obtained by the use of Kirchoff's Laws:

$$E + I_3 R_p = u E_1 \quad (1)$$

$$I_4 = j w c_1 E \quad (2)$$

$$I_3 = I_4 + I_1 + I_m \quad (3)$$

$$I_5 = I_2 + I_m \quad (4)$$

$$E + j w M I_2 = Z_1 I_1 \quad (5)$$

$$j w M I_1 \quad Z_2 I_2 + I_5 Z_g \quad (6)$$

$$I_1 z_1 + I_2 z_2 = j w M I_2 + j w M I_1 + \frac{I_m}{j w C_m} \quad (7)$$

$$I_1 Z_1 - j w M I_2 - E = 0 \quad (8)$$

$$-j w M I_1 + Z_2 I_2 + (I_2 + I_m) Z_g - = 0 \quad (9)$$

$$I_1 (Z_1 - j w M) + I_2 (Z_2 - j w M) - \frac{I_m}{j w c m} = 0 \quad (10)$$

On solving these last three equations for I_1, I_2, and I_3 we get:

$$I_1 = \frac{E}{D} \frac{Z_2}{j w C_m} + \frac{Z_g}{j w C_m} + Z_2 Z_g - j w M Z_g, \text{ where} \quad (11)$$

$$D = \left(\frac{Z_1 Z_2 + W^2 M^2 + Z_1 Z_g}{j w C_m}\right) + Z_g (Z_1 Z_2 + W^2 M^2) \quad (12)$$

$$I_2 = \frac{E}{D} \left(\frac{W M}{W C_m} + Z_g Z_1 - j w M Z_g\right) \quad (13)$$

$$I_m = \frac{E}{D} (W^2 M^2 + Z_1 Z_2 + Z_1 Z_g - j w M Z_g) \quad (14)$$

$$I_3 = I_4 + I_1 + I_m \quad (3)$$

$$I_4 = j w C_1 E \quad (2)$$

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$$I_3 = j w C_1 E + \frac{E}{D}\left(\frac{Z_2}{j w C_m} + \frac{Z_g}{j w C_m} + Z_2 Z_g - 2 j w M Z_g + W^2 M^2 + Z_1 Z_2 + Z_1 Z_g\right) \quad (15)$$

$$I_3 = j w C_1 E + \frac{EP}{D} = \frac{HE}{D} \quad (15a)$$

where $$P = \frac{Z_2}{j w C_m} + \frac{Z_g}{j w C_m} + Z_2 Z_g - 2 j w M Z_g + W^2 M^2 + Z_1 Z_2 + Z_1 Z_g \quad (16)$$

and $$H = j w C_1 D + P \quad (17)$$

$$E + I_3 R_p = u E_1 \quad (18)$$

$$E = \frac{u E_1 D}{D + H R_p} \quad (19)$$

but $$E_g = I_5 Z_g = (I_2 + I_m) Z_g \quad (20)$$

$$E_g = \frac{E}{D}\left(\frac{W M}{W C_m} + W^2 M^2 + Z_1 Z_2\right) Z_g \quad (21)$$

Amplification =

$$\frac{E_g}{E_1} = \frac{u Z_g}{D + H R_p}\left(\frac{W M}{W C_m} + W^2 M^2 + Z_1 Z_2\right) \quad (22)$$

This expression is rather involved and requires a certain amount of tedious work in using it, but it will give the correct amplification for the assumed conditions.

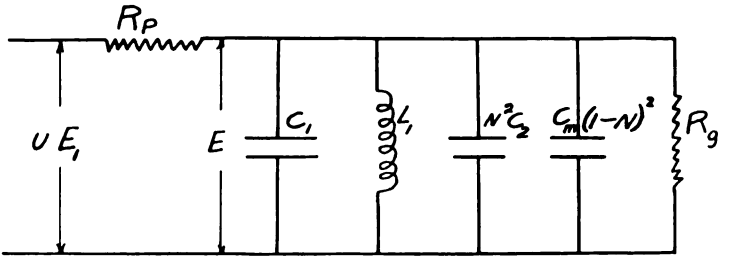

Figure 3

To simplify matters the following assumptions will be made:

Let $R_1 = 0$

$R_2 = 0$

This will not cause any serious error as R_p which is in series

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with R_1 is usually very large. Similarly R_2 is very small as compared with R_g or $\frac{1}{j w C_2}$

$$M = \sqrt{L_1 L_2} \qquad L_2 = N^2 L_1$$

This assumption that $M = \sqrt{L_1 L_2}$ means that the coupling coefficient between L_1 and L_2 is unity, while the leakage in a good transformer will be small (1 per cent or less), the form of the amplification curve at the higher frequencies is largely controlled by the leakage inductance. The amplification obtained by the use of this assumption will therefore not be correct at the high frequencies. Fig. 4 shows the amplification curves obtained by the exact and approximate

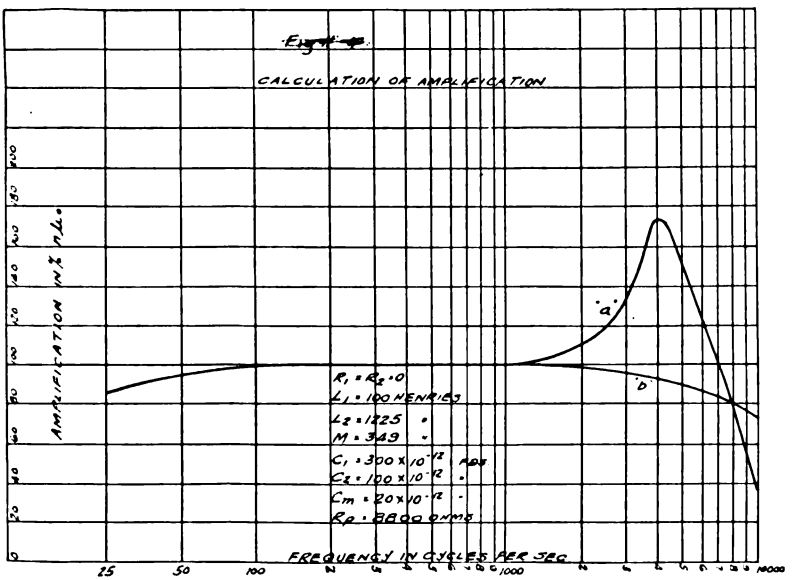

Figure 4—Calculation of Amplification.

methods. The only disagreement is in the upper range. Curve "a" is exact and "b" an approximation.

Since $M = \sqrt{L_1 L_2}$ and $L_2 = N^2 L_1$

$$I_s = j w c_1 E + \frac{\frac{E}{R_g}}{N^2} + j w C_2 N^2 E + \frac{E}{j w L_1}$$

$$+ j w c_m (1 - N)^2 E \text{ and } I_5 = E \left(\frac{N}{Z_g}\right) \qquad (23)$$

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On examining the equation for I_3 it will be seen that the circuit in Fig. 3 will be equivalent to it.

Let $r_e + j x_e$ be the equivalent impedance of the transformer,

then $$I_3 = \frac{E}{r_e + j x_e} \tag{24}$$

$$E = u E_1 - I_3 R_p = u E_1 - \frac{E R_p}{r_e + j x_e} \tag{25}$$

$$E = \frac{u E_1 (r_e + j x_e)}{R_p + r_e + j x_e} \tag{26}$$

$$\begin{aligned} E_g = I_5 Z_g = \frac{E N}{Z_g} \times Z_g = N E \\ = \frac{N u E_1 (r_e + j x_e)}{R_p + r_e + j x_e} \end{aligned} \tag{27}$$

$$\text{Amplification} = \left| \frac{E_g}{E_1} \right| = \left| \frac{n u E_1 (r + j x)}{(R_p + r_e + j x_e) E_1} \right|$$

$$= \left| \frac{N u (r_e + j x_e)}{R_p + r_e + j x_e} \right| = N u \frac{\sqrt{r_e^2 + x_e^2}}{\sqrt{(R_p + r_e)^2 + x_e^2}} \tag{28}$$

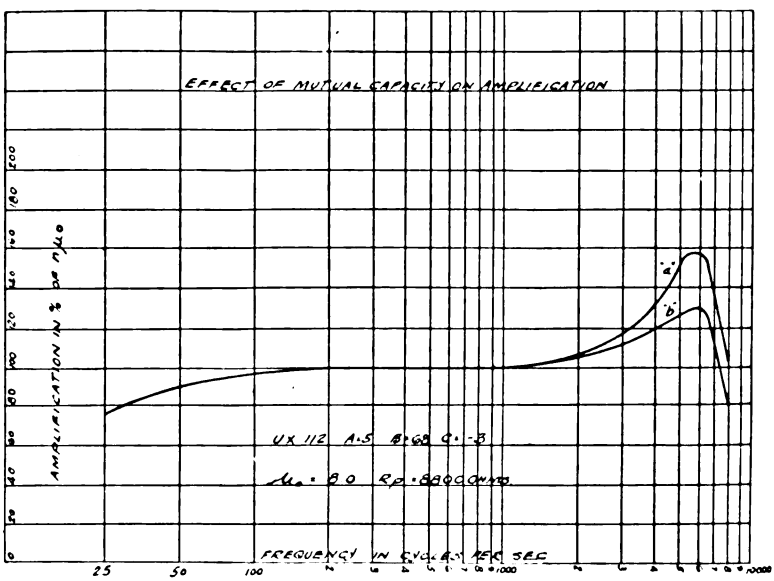

Figure 5—Effect of Mutual Capacity on Amplification.

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The greater $r_e + j x_e$ is the closer the amplification approaches to $n u$.

For maximum amplification $r_e^2 + x_e^2$ must be very large as compared with $R_p^2 + 2 r_e R_p$.

r_e is usually small and if the grid filament resistance is large enough to be neglected r_e will be zero.

The only characteristics of the tube required to give the amplification curve is the plate impedance and the amplification factor of the tube. There is no simple relation between these two factors which would give us a means of comparing the relative merits of tubes.

If the direction of the secondary current is reversed,

$$P = \frac{Z_2}{j w c m} + \frac{g}{j w c m} + (Z_2 + 2 j w M + Z_1) Z_g + W^2 M^2 + Z_1 Z_2$$

The rest of the proof is similar to the first part.

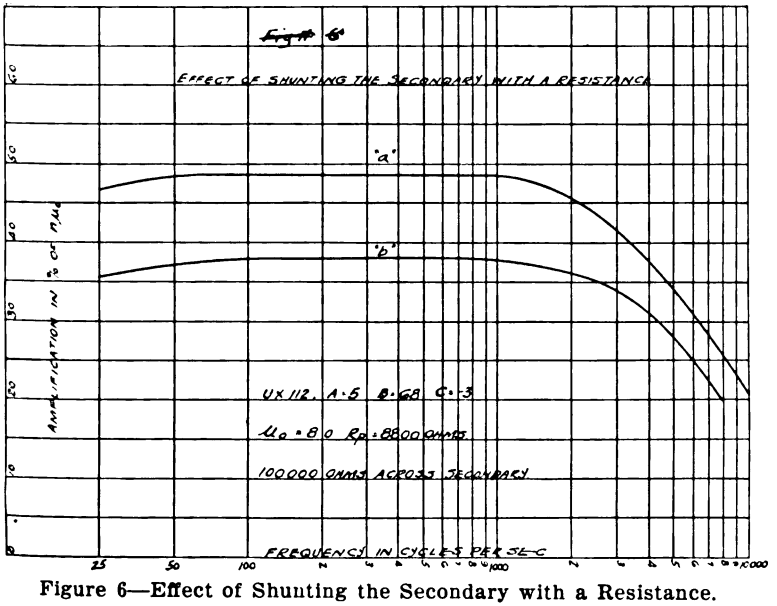

Figure 6—Effect of Shunting the Secondary with a Resistance.

When $R_1 = R_2 = 0$ and $M = \sqrt{L_1 L_2}$
the same equivalent circuit will be obtained with the capacity C_m now increased to $(1 + N)^2$ instead of $(1 - N)^2$.

Fig. 5 shows the effect of the mutual capacity on the amplifications curve of the transformer. Curve "a" was

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obtained with the mutual capacity of the transformer neglected. Curve "b" shows the effect of the mutual capacity. A, B and C are filament, plate and grid battery voltages.

The curves in Fig. 6 show the effect of a shunted secondary. Curve "a" was calculated from assumed constants while curve "b" is a curve obtained from test. The form of the two curves is very similar.

In conclusion the results obtained for the assumed conditions are as follows:

The maximum amplification is obtained when the equivalent impedance of the transformer is very large as compared with the plate impedance of the tube. There is no simple relation between the plate impedance and the amplification constant of the tube, which will give the performance of the tube when used with audio frequency transformers. The mutual and self capacity of the coils acts as a pass at the higher frequencies and the effect depends on the turn ratio of the transformer.

SUMMARY: In the foregoing paper a method for calculating the amplification curve of an audio frequency transformer is developed in terms of the usual constants of the transformer and of the tube. The distributed capacity of the coils and the mutual capacity between the primary and secondary coils are represented by lumped capacities. The exciting current of the transformer is neglected. As the equation for the amplification in the vector form is rather involved an approximate formula is then developed and its limitations pointed out.

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NOTES ON THE TESTING OF AUDIO FREQUENCY AMPLIFIERS*

BY
EDWARD T. DICKEY

Fundamentally, the testing of audio frequency amplifiers consists of an examination of the operational and structural features of the amplifier with the view to determining the following points:

1. The voltage amplification, with special regard to the relative amplification at various points along the audio frequency range.

2. Ability to produce the necessary output power for the particular use for which it is intended, without distortion of the impressed wave form.

3. General operational and structural practicality with reference to such points as: freedom from audio frequency howling, susceptibility to howling through lengthening of the input and output leads, correct values of grid and plate potential for the particular tubes used, etc.

Of these points, the most important, and those requiring the most equipment and care in their determination, are undoubtedly the first two. The greater portion of this paper, therefore, will be devoted to consideration of these points.

It is, of course, understood that a test of the audio frequency amplifying system constitutes by no means the entire story on the fidelity of a complete receiving set. The various tuning, amplifying, and detecting stages preceding the audio frequency amplifier, all have their effect on fidelity. Those interested in a discussion of the other factors affecting fidelity and their measurement, are referred to a contemporary paper on "Notes on Receiver Measurements" by Messrs. T. A. Smith and G. Rodwin.

1. Voltage Amplification Measurement

It is not intended in the short space of this paper, to discuss all the methods, or even all of the most important

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methods, of attacking this problem. Discussion will be confined to the methods which have been found by the author to give the best results in combined accuracy, speed of test, and simplicity of operation.

The apparatus necessary consists of the following three essentials:

A. A source of variable audio frequency potential, capable of supplying the required amplifier input potential, and producing oscillations whose wave form differs from the sinusoidal to a negligible degree.

B. Means for accurately measuring the potential impressed upon the input of the audio frequency amplifier under test.

C. Means for accurately measuring the output potential from the amplifier, without appreciably influencing any characteristic of the amplifier.

Description of the apparatus considered most suitable for these three purposes is given below:

A. *The Oscillator.* The source of audio frequency potential found most satisfactory was an oscillator of the beat frequency variety, employing the difference frequency of two radio frequency oscillators. By combining and rectifying these frequencies, and then amplifying the resultant frequency, it was possible to obtain from the oscillator of the type used by the author, audio frequency potentials of the order of 50 volts. The wave shape of the oscillations obtained from a properly constructed oscillator of this type, can be made to contain not more than 1 or 2 per cent of harmonics, and the output potential will remain effectively constant over the frequency range from 30 to 10,000 cycles approximately, without readjustment.

It has become a fairly well established practice in radio laboratories, to use a logarithmic scale for the frequency, in plotting audio frequency characteristic graphs. By properly shaping the plates of the frequency adjusting variable condenser, it is possible to make the frequency produced by the oscillator vary logarithmically with respect to the condenser dial. The advantage of this will be evident when the method of using this oscillator in testing audio frequency amplifiers is described. (See section on General Description of Amplification Test Method).

B. *Input Potential Measurement.* The output of the

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oscillator used is adjustable internally, and thus for this measurement, it is only necessary to employ a thermo-milliammeter of reliable make, which is accurate over the audio frequency range. Associating this meter with the necessary series resistance, which must be accurate over the range of audio frequencies used, potentials over the desired range may be measured and applied to the amplifier input. As stated previously, the output potential of the oscillator when once set, will remain appreciably constant over the range from 30 to 10,000 cycles, which is sufficient for the usual amplifier measurements. It is considered good practice to read the input voltage again at the end of the test run, to make sure that the oscillator is working properly.

C. *Output Potential Measurement.* Considering the requirement that the measuring device used here shall not affect the characteristics or output of the amplifier, the most obvious instrument to use is a voltmeter of the vacuum

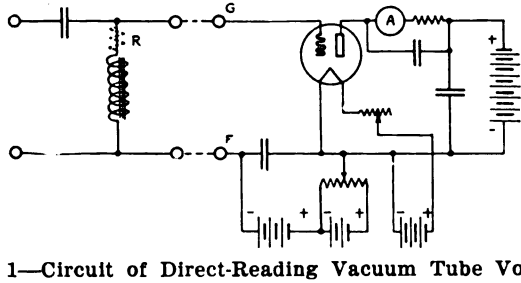

Figure 1—Circuit of Direct-Reading Vacuum Tube Voltmeter

tube type. To simplify the operation, and to permit readings to be taken rapidly, a direct-reading type of vacuum tube voltmeter is desirable.

To facilitate operation in connection with the curve drawing mechanism, which will be described later, a tube voltmeter having very nearly a linear characteristic is needed. By the use of a UX-171 tube, it was found possible to construct a voltmeter having a number of very desirable features for amplifier output measurement purposes. This tube has a low plate impedance, and a high grid bias potential, and it is possible, by its use, to obtain a tube voltmeter which is capable of measuring directly potentials up to approximately 70 volts, and which has a reasonably straight

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characteristic curve from about three volts to the end of the scale. A high plate potential, a high series plate resistance, and a microammeter having a full scale deflection of 100 microamperes, were contributing factors in producing the straight line characteristic.

The circuit used is shown in Fig. 1. Because of the relatively low plate current (maximum approximately 100 microamperes), it is possible to run the filament of the tube below its full rated value (4.5 volts permissible), thus prolonging its life and constancy of calibration. For increasing the accuracy with which the zero reading of the meter might be set, it was found best to use a reading of 10 microamperes as zero, calibrating the meter scale from that point. The tube voltmeter is calibrated using the oscillator and thermoammeter described before. On turning on the filament and grid potentials, it is merely necessary to make a

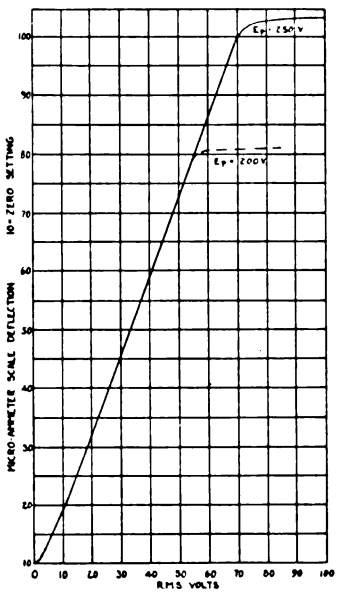

Figure 2—Calibration Curve of Direct-Reading Vacuum Tube Voltmeter (0-70 v. Range).

slight adjustment of the potentiometer until the microammeter reads ten. Changes in the zero reading during operation are relatively negligible. The condensers connected across various points of the circuit are used to pre-

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vent inequalities in reading at the high and low frequencies. As shown, the instrument is equally accurate over the entire audio frequency range within observational error percentage.

At first glance it seems somewhat dangerous to use a plate meter of such high sensitivity in a circuit having such a high applied potential. By a very simple calculation, however, it will be evident that even if the plate and filament of the tube should become short circuited, the high value of series plate resistance would prevent the plate meter from being overloaded. As will be seen from the calibration curve (Fig. 2) the plate meter reading remains practically constant after it has reached a reading of approximately 104 microamperes. While slightly past the end of the scale, this will not injure the instrument.

In amplifiers where there is no direct current flowing through that portion of the output circuit across which the tube voltmeter is connected, the binding posts marked *G* and *F* are connected direct to the amplifier circuit. In cases where the amplifier output circuit contains a direct current component, use is made of the condenser and choke combination shown to the left of the circuit in Fig. 1. An inductance is used for the grid potential return, rather than an ohmic resistance, because of the trouble experienced with the condenser and resistance leak combination, due to the slowness of such a system in returning to normal, after a sudden high potential surge such as is often experienced through induction effects on sets operating on commercial lighting circuits. The resonant period of the inductor and capacitator combination must be outside of the range of audio frequencies used. Actual test has shown this circuit to give appreciable equality of operation over the usual audio range, if the impedance of the choke is kept at least ten times as great as that of the output device. This is quite easy if a loud speaker of usual type is connected to the amplifier output. In cases where the output potential is to be measured across a relatively high ohmic resistance, it is advisable to use an ohmic resistance leak in series with the choke as shown at *R* in Fig. 1. This resistor should have a value approximately 20 times as great as the resistance across which the potential is to be measured.

By a simple alteration of circuit constants, the tube volt-

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meter may be arranged for the measurement of potentials of the order of a few volts. In Fig. 3 is shown a circuit making available two ranges by the mere throwing of a double-pole double-throw switch. While the low range will not have quite as straight a characteristic as the higher range, it is still quite satisfactory, as will be seen from Fig. 4, which gives the curve for a range having a maximum of about seven volts. A very compact and convenient tube voltmeter

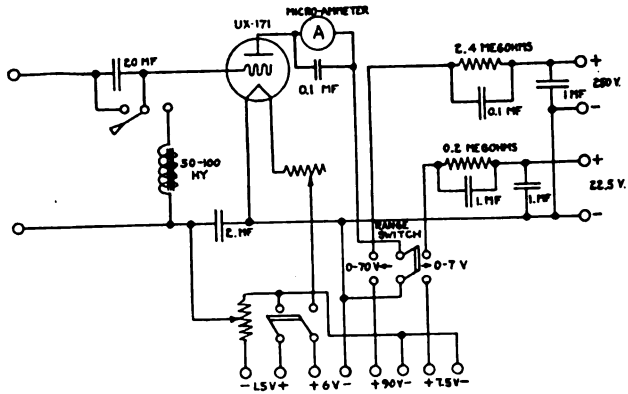

Figure 3—Circuit of Direct Reading Vacuum Tube Voltmeter Arranged for Measurement of Two Ranges of Potentials (0-7 v. and 0-70 v.)

of the type described above is shown in Fig. 5.

Like all vacuum tube voltmeters using a d-c. indicating instrument, the one described here will have its accuracy affected by the wave form of the potential under measurement. It will be found to be considerably better in this regard than some tube voltmeters, however. Those interested in the effect of wave form on the accuracy of various types of tube voltmeters, are referred to two recent articles on "The Thermionic Voltmeter" by Messrs. W. B. Medlam, and U. A. Oschwald, in the October and November, 1926 issues of *Experimental Wireless and the Wireless Engineer.*

If calibrated by the use of a potential of good wave shape, this tube voltmeter should retain an accuracy of approximately 1 per cent of full scale for a considerable period. Briefly, the points upon which retention of its accuracy depends are, in order of importance:

1. The accuracy of the plate meter, (directly dependent.)

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2. Retention of the same value of series plate resistance. This value can be checked quite simply, however. By merely connecting the plate terminal of the tube socket to the negative filament terminal, and lowering the plate potential somewhat, the resistance can be directly obtained by calculation from the impressed potential and plate meter reading.

3. Tube constants. If a tube with normal constants is selected, the low plate current and filament current result in retention of appreciably the same constants for a considerable period.

4. Filament potential. Variation of 10 per cent causes a variation of less than 1 per cent in reading, and therefore accuracy is relatively independent of this factor within these limits.

5. "Zero setting" of the plate meter. Setting for 10

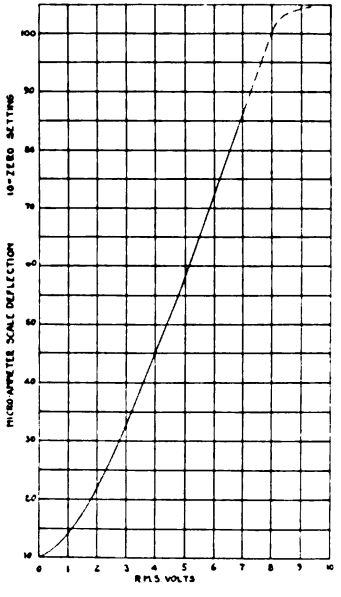

Figure 4—Calibration Curve of Direct-Reading Vacuum Tube Voltmeter (0-7 v. Range)

microamperes is important, but is extremely easy to check and, when set, practically never varies over ½ per cent even over a considerable period of continued use.

6. Plate potential. As long as the zero setting is correct,

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a variation of as much as 20 per cent in plate potential will cause less than 1 per cent variation in reading, up to the "cut off" point. This point, i. e., the point above which the plate current will not rise with increased input potential, will be lowered. This can be noted in Fig. 2, where the horizontal line marked "200 volts" shows where this point would be for this value of plate potential.

7. Grid potential. It is not necessary to know the actual value, since this is set by watching the plate meter. Its numerical value is of interest only when one wishes to be sure that the tube is not acting as a load across the circuit under measurement, by drawing grid current.

Test of Multi-Stage A. F. Amplifier

It is important to arrange test conditions so as to approach, as nearly as possible, those which will exist in actual operation. In this connection, the following points are important:

1. Input and output impedances should be the same as in actual use.

2. Long input or output leads, or anything tending to produce interstage couplings which would not be present in normal operation, should be avoided.

3. Values of input potential should be chosen so as to avoid overloading the output tube, and tests should be made preferably using two or more values of input potential, to discover any alteration in operating characteristics with varying inputs.

4. If the amplifier input consists of a transformer winding, provision should be made to have normal direct current flowing through the winding, if the latter is to be connected directly to a tube plate circuit, as is usually the case.

5. Conditions of grounding or shielding should be similar to those of actual use.

In testing a complete amplifier, the types of tubes which are intended to be used in it are generally known. The only points which may be in doubt are usually the characteristics of the output device (loud speaker), and the tube (generally detector) which will precede the amplifier. Usually the amplifier input will be in the form of a transformer primary winding, in which case amplification curves may be taken

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with several values of series input resistance, and primary direct current.

A simple circuit for use in obtaining the amplification characteristic is shown in Fig. 6. The adjustable resistance R_2 provides means for simulating the plate resistance of the tube preceding the amplifier. R_1 should be small in value compared wih the impedance of the thermometer and os-

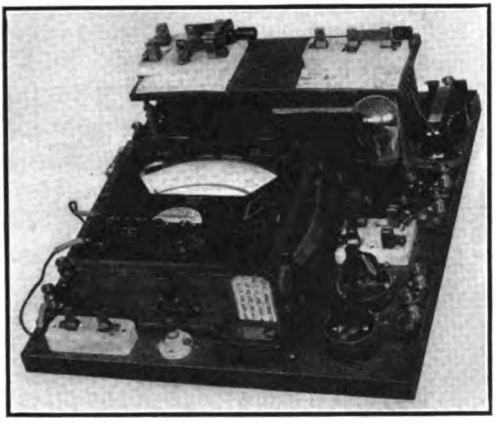

Figure 5—Vacuum Tube Voltmeter.

cillator output circuit. In case R_1 is appreciable with respect to R_2, the sum of the two should be made equal to the desired series input resistance. By means of a source of adjustable potential E_1 and a milli-ammeter, a direct current, equal to the plate current of the tube feeding the input, may be set up in the input primary winding.

If the amplifier input is a tube grid circuit, the oscillator potentials may be impressed directly thereon. In the case of resistance coupled amplifiers, the input will probably be in the form of a plate resistance. Various series resistances to imitate tube plate resistance, may be tried between this and the input potential source, but it will generally be unnecessary to put a direct current through the plate resistance. If there is doubt as to the ability of this resistance to carry the necessary current quietly, and without changing its resistance, it is advisable to investigate the resistance in question, separately.

In cases where the amplifier input circuit is not fed from

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a tube, electrical characteristics simulating as nearly as possible those of actual use, should be arranged at the point where the input potential is impressed on the amplifier.

In case the characteristics of the loud speaker with which the amplifier is to be used are not known, it is generally best to use a resistance in the output circuit of the value for which the output tube is designed. From the

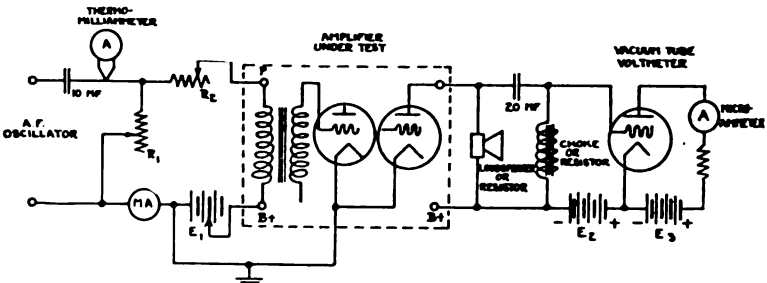

Figure 6—Test Circuit for Transformer-Coupled Amplifier of Several Stages.

curve thus obtained, and a knowledge of the plate impedance of the output tube, it is possible to determine by calculation, the characteristics of the amplifier with any loud speaker whose impedance characteristics are known.

Testing Individual Amplifying Transformers

While it is the primary purpose of this paper to discuss the testing of complete amplifiers, the testing of individual audio frequency transformers is believed to be of sufficient importance to warrant a brief separate treatment here. This is a more specialized problem than the testing of completed amplifiers, but is of particular importance for design purposes.

It should be noted that it is practically impossible to construct a test set up which will simulate accurately the circuit reactions to which a transformer will be subjected in a completed amplifier. Thus it is recommended that not too much dependence be placed on results obtained from test of a single transformer, which is intended for use in a multistage amplifier. Features of interstage coupling, input and output coupling, tubes, transformers in other stages, and loud speaker characteristics, all enter into the problem, mak-

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ing it exceedingly difficult to duplicate in the test of a single transformer, the conditions under which it will work in the completed amplifier.

In Fig. 7 will be found a circuit which can be made to simulate to some extent, the characteristics of a single stage of a multi-stage amplifier.

The remarks made previously in connection with Fig. 6 regarding the input circuit adjustments, apply here also. It is recommended that the circuit be grounded at the points indicated, since these points will, in general, be at ground potential in the final amplifier. If the core of the transformer is to be grounded in the final amplifier, it should be grounded during the test. It is important that the grid potential battery of the output measuring tube voltmeter be connected at the point indicated, and not in the grid lead as is sometimes done, otherwise excessive capacity will exist across the transformer secondary. If the grid-filament capacitance of the tube-voltmeter tube is not sufficiently close to that of the tube which is to be used on the secondary side of the transformer, either of two possible expedients may be used:

1. An extra frequency run may be made with a small condenser C connected across the transformer secondary. From the two curves obtained, and knowing the grid-filament capacitance of the tube voltmeter, and of the tube which is to be used with the transformer, it will be possible to determine the curve for the tube to be used by extrapolation.

2. Use the type of tube which will be fed from the secondary of this transformer in the final amplifier, and connect the tube voltmeter across a resistance in the plate circuit of this tube. Tested in this way, of course, the voltage amplification measured includes the amplification factor of the tube following the transformer. To get the voltage amplification of the transformer, it is therefore necessary to divide the values obtained by the amplification factor of this tube.

If it is desired to determine the general characteristics of a transformer, so that its behavior with various types of tubes may be deduced, this may be done with fair accuracy by calculations based upon certain measurements. To do this it will be necessary to find:

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1. The curves of reactance and effective resistance variation with frequency of the primary winding, with the secondary open. These must be determined for two or more values of secondary shunting capacitance, within usual or desired range of filament to grid tube capacitance, and with several

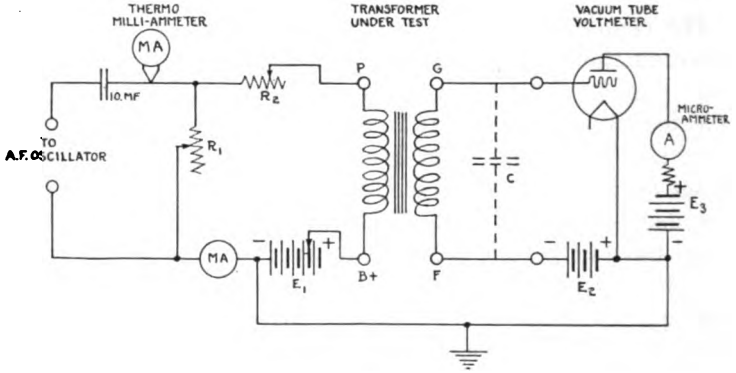

Figure 7—Test Circuit for Single Interstage Transformer.

values of direct current through the primary winding. Values of direct current (simulating tube plate current) as high as the greatest which it is proposed to use, should be tried.

2. The ratio of potentials across primary and secondary windings should be obtained for the entire frequency range. Here also it will be necessary to take measurements with various values of secondary capacitance, and primary direct current, and curves for the complete frequency range should be plotted.

Then letting e_1 = Potential generated in plate circuit of the tube connected to primary winding,

e_2 = Potential across the secondary, as result of e_1 in tube,

r = Plate resistance of the tube,

R = Effective resistance of Transformer Primary,

X = Reactance of Transformer Primary,

E_2/E_1 = Ratio of potentials measured directly across Secondary and Primary respectively.

The following formula gives the relation between these factors:

$$e_2 = \left(\frac{E_2}{E_1}\right) \frac{e_1}{\sqrt{(r+R)^2 + X^2}} \sqrt{R^2 + X^2}$$

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Thus, with the aid of the curves, and this formula, it is possible to predict with fair accuracy the action of a transformer with any tubes, knowing the plate resistance and plate current of the preceding tube, and the grid-filament capacitance of the tube following.

Two suggested methods of making the measurement of

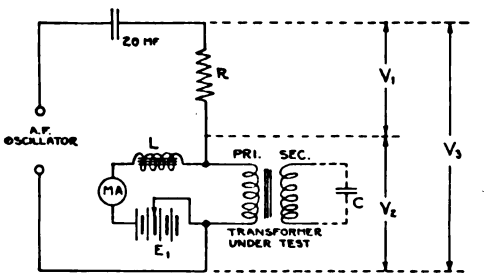

Figure 8—Voltmeter Method for Determining Reactance and Effective Resistance of Transformer Primary.

the reactance and effective resistance of the transformer primary winding are given below:

1. By means of the ordinary inductance bridge, arranged preferably with both the inductance and resistance values of its standards continuously variable.

2. It is often desired to perform this measurement with a definite value of a-c. potential applied to the primary. In this case, the well known three-voltmeter method can be used to advantage. Fig. 8 shows in outline the circuit, with means for applying a d-c. potential to the primary. The reactance of the choke L must be sufficiently high with respect to that of the transformer primary to make the effect of the shunting circuit negligible, or its inductance value must be known and its effect allowed for in the calculations.

The use of the tube voltmeter is recommended for obtaining the values of V_1, V_2, and V_3. The calculations involved in getting the values of primary reactance and effective resistance from these potentials is well known, and for the sake of brevity will not be repeated.

General Description of Amplification Test Method

After the apparatus and device under test is set up, and necessary adjustments to bring about normal operating con-

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ditions are made as described in connection with Figs. 6 and 7, there remains the choice of two methods of taking the amplification versus frequency characteristic.

1. The oscillator may be set at various frequencies throughout the desired range, and a reading of the tube voltmeter taken for each frequency. From the data obtained, a curve of the amplifier characteristic may be plotted. This may be plotted either directly in output volts, or, by dividing the output potentials by the constant input potential, a curve showing the voltage amplification at each frequency can be plotted. The closer together the points have been taken, the more nearly will the curve show the operation at all frequencies.

2. A quicker method, and one in which a continuous measurement is made of all points along the frequency range is shown in general outline, with the necessary apparatus arrangement in Fig. 9. Here the shaft of the

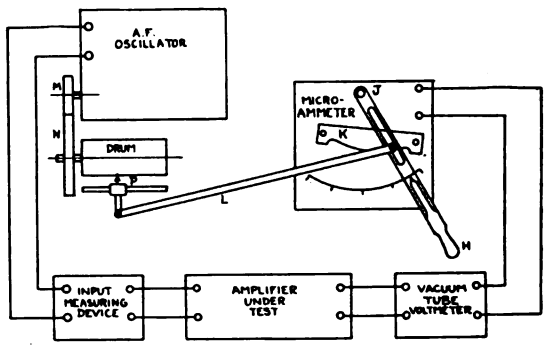

Figure 9—Audio Frequency Amplifier Test Set Up.

oscillator variable condenser is geared, by some convenient means, shown here as the gears *M* and *N*, to a drum which is of the proper size to hold a standard sheet of plotting paper. Paper having a logarithmic scale for frequency, and linear scale for amplification is used. The ratio of the gears *M* and *N* is such, that the frequency lines on the plotting paper which come under the pen *P* as the condenser shaft is rotated, may be made to correspond with the frequency produced by the oscillator at the moment. If then the distance of the pen P above the zero axis of the paper, is made to correspond to the output voltage reading of the amplifier, we

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have a means for drawing the amplifier characteristic curve directly as the test is being made.

By means of the arm and pointer *H*, which is pivoted at *J*, it is possible to follow the movements of the tube voltmeter microammeter needle. The movements of the arm *H* are communicated to the pen *P* through the lever *L*, and by proper shaping of the guide template *K*, it is possible to have the voltages measured by the tube voltmeter correspond to the deflection of the pen *P* in terms of the ordinate divisions of the plotting paper. Since the templates for the 70-volt and 7-volt ranges differ somewhat in shape, they should be so constructed as to be removable and interchangeable within the curve drawing mechanism. Fig. 10 shows the drum, microammeter, pen, and attendant curve drawing mechanism. Fig. 11 shows this equipment attached to the beat frequency audio oscillator.

If desired, the oscillator condenser may be rotated slowly by motor or clockwork, and the observer needs only to follow

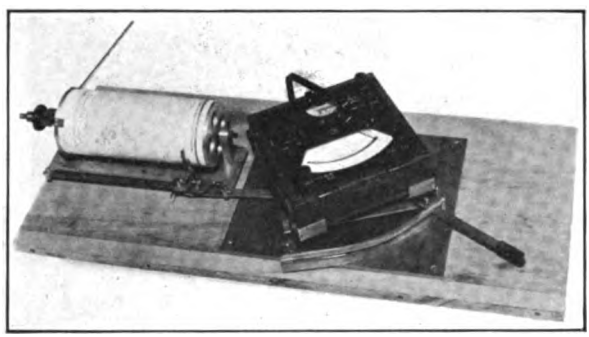

Figure 10—Curve Drawing Mechanism.

the movements of the microammeter needle carefully with the pointer *H*. Upon reaching the upper end of the frequency range, the condenser drive is disconnected, and the paper removed from the drum. Upon the paper, the curve of output voltage vs. frequency for the amplifier under test will be found.

2. Test for Wave Form Distortion

Returning now to the second of the tests which are necessary in examining the operation of an audio frequency

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amplifier, i. e., test of its ability to handle the necessary output energy without distortion of the impressed wave form, we find that test of this point includes determination of the input potential which will cause overloading distortion—and thus the permissible input potential—and for accurate data, involves determination of the percentage of harmonics which are added to the original frequency as it passes through the amplifier.

The principal points in a transformer coupled amplifier where wave shape distortion may occur are, in the magnetic circuit of the transformers, due to excessive flux density, and in the tubes. If the transformers are well designed, the first point can be neglected. Distortion in the tubes will generally be caused by the non-linearity of the grid potential-plate current characteristic, or by the grids of one or more tubes becoming positive with respect to their fila-

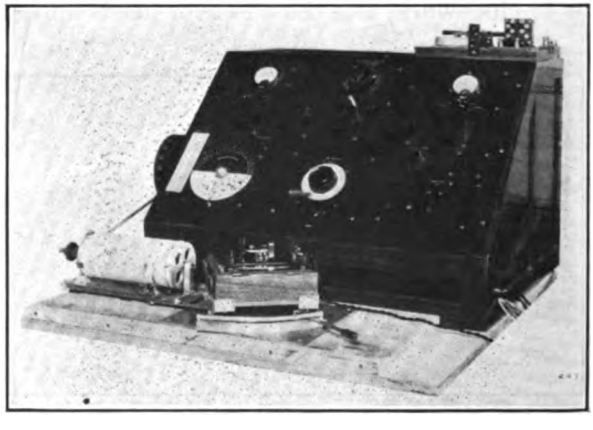

Figure 11—Curve Drawing Mechanism Attached to A. F. Oscillator.

ments. On high input potentials, one or more of the grids will become positive, and the resulting grid current and lowered grid filament impedance of the tube, will cause very serious distortion. This type of distortion is by far the more serious of the two caused by the tubes, and is probably the only type which would be noticed by the average person listening to the loud speaker output from the amplifier.

It is often desirable to test an amplifier only for this more obvious type of distortion, and such a test is quite

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simple. The tube which is generally the first to show signs of grid current is the one in the last stage. By putting a low reading d-c. milliammeter or microammeter in series with the grid of this tube, and gradually increasing the input potential, the maximum permissible input will be just below that value which causes the meter to read in a direction showing that the grid is positive. (Most tubes have a small constant grid current in the opposite direction, which should be disregarded). It is advisable to make the test at several different audio frequencies, and if there is reason to believe that the grids of any of the other tubes in the amplifier are becoming positive, they should be tested in the same way.

For proper and safe operation, the amplifier should be capable of giving more output signal voltage than is

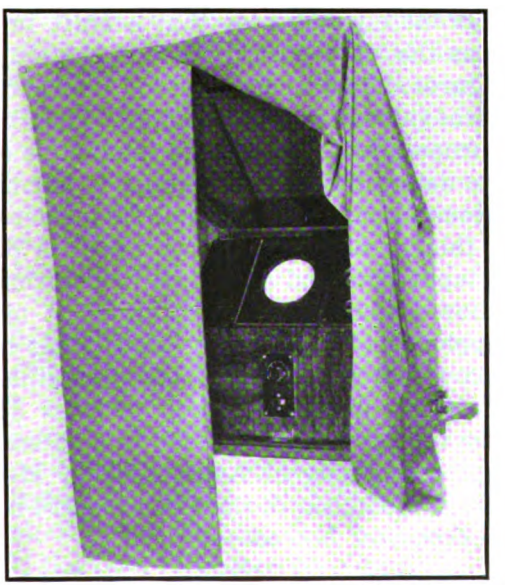

Figure 12—Cathode Ray Oscillograph Tube Mounting.

normally required, before any of its tubes show signs of grid current at any point of the audio frequency range.

For those who prefer a method giving more accurate numerical indication of the amount of wave shape distortion present, analysis of the output wave by means of the cathode

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ray oscillograph is recommended. If potential from the source feeding the input of the amplifier under test, is impressed across the "horizontal motion" plates of the oscillograph, and potential from the amplifier output is impressed across the other set of plates, a figure will be traced on the oscillograph screen, which will show, by its shape, with what fidelity the amplifier reproduces the wave shape impressed upon its input. Fig. 12 shows the mounting used by the author for the cathode-ray oscillograph tube. To facilitate tracing the curves which appear on the screen, a hood was built over the tube mounting cabinet. One side of the hood cloth is shown drawn aside to permit a view of the tube.

If a single straight line is traced, perfect fidelity and no phase shift is indicated. In cases where a phase shift occurs in the amplifier, the figure will assume an elliptical instead of a linear shape. This makes analysis of the figure more difficult, but the elliptical shape is no sign of wave distortion. It merely indicates that a shift in phase has taken place as the wave passed through the amplifier. Oscillograph figures should be obtained for several frequencies, and for several values of input potential. Then by an application of harmonic wave analysis to the figures obtained, it is possible to determine the percentage of harmonics present, for various input potentials and frequencies. Depending on the fidelity of output wave which is desired, the permissible input potential may then be determined.

It is believed that an oscillographic test of this kind should be applied more frequently than it has been in the past, to the testing of audio frequency amplifiers. As the scope of this paper prevents further or more detailed discussion, those interested are referred to a recent article on "Load Carrying Capacity of Amplifiers" by Messrs. F. C. Willis and L. E. Melhuish, in the October 1926 issue of *The Bell System Technical Journal.*

3. General Operation

The general points for test and examination mentioned under the third section are usually different for each amplifier, and no very definite rules for investigation procedure can be given. The method of investigation depends very largely on the particular amplifier, and no general rules can

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be laid down. The experience and ingenuity of the test engineer must be relied upon generally to determine the method of test. Little if any definite apparatus can be recommended since these tests require, in many cases, merely intelligent and careful inspection. One important point may be mentioned in this connection, however, i. e., in regard to checking whether the proper plate potential is being used on the tubes in a resistance coupled amplifier. The plate potential may be measured with an electrostatic voltmeter at the plate of each tube while the amplifier is in operation, or else its value may be calculated from measurement of the plate current and plate coupling resistance value for each tube.

No attempt has been made in this paper to give anything but the essentials of audio amplifier testing. It will usually be found in practice that each amplifier constitutes to a greater or lesser extent a special problem, since each will differ from the others in certain points. The essential features to be tested for, together with a brief description of the apparatus found most suitable for making the tests, have been given. There are many other methods of taking amplifier characteristics, but it is believed that the methods described here are as good as most of the other methods, and perhaps excel some in points of accuracy, convenience, and speed of measurement.

In conclusion, the author wishes to acknowledge his indebtedness to the following members of the staff of this laboratory:—Dr. Walter Van B. Roberts for valuable suggestions regarding the method of determining the general characteristics of individual amplifying transformers by combined measurement and calculation; Dr. Irving Wolf for the ideas embodied in the method of directly drawing amplifier characteristic curves described in this paper; and Mr. David Grelich for his assistance in connection with the development of the tube voltmeter described.

SUMMARY: The more necessary of the various points which must be tested in examining the performance of audio frequency amplifiers are outlined, and a method of test procedure, found by the author to have desirable characteristics from the points of view of accuracy, speed, and simplicity of operation, is described.

The method used permits the complete curve of amplifi-

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cation vs. frequency for the amplifier under test, to be drawn directly by the test equipment in a very short time. A type of tube voltmeter found convenient for measurement of amplifier output potential is described. Brief discussion is given of the testing of individual audio frequency amplifying transformers, and some test methods applicable thereto are suggested. Amplifier wave form distortion, and overloading are discussed, and methods for the test thereof are recommended.

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VARIATIONS IN HIGH-FREQUENCY GROUND WAVE RANGES

By
A. H. TAYLOR
(Naval Research Laboratory)

The following information obtained from experimental work made by two naval ships on high frequency ground wave ranges is of interest because it indicates unexpected differences between day and night values, which are not at present predictable from theoretical consideration.

The ground wave of a high-frequency station is usually recognized readily enough if observations are taken well within the normal skip distance area because the ground wave is normally quite steady; that is, free from fading. The observations herein reported were taken on frequencies in the 12,000 and 16,000-kc. bands at distances between zero and 53 miles. The power of the transmitter was approximately 100 watts in the radio frequency circuits. Probably 60 per cent of this power went into the antenna. In agreement with earlier observation it was found that ground waves traveled further over the sea than over the land, so that with daylight conditions on a transmitter of considerable power it was not easy to locate sharply the skip distance which made itself felt rather as a depression in the curve of audibility of the function of the distance rather than an absolute absence of signal, unless frequencies higher than 10,000 kc. were used.

The daylight ground wave ranges of these transmitters working in the 12,000-kc. band were approximately 12 miles, whereas the same transmitters had a satisfactory night range in the same frequency band of something somewhat in excess of 50 miles. The extreme limit of audibility was estimated roughly to be about 75 miles. Actual observations were taken at 53 miles. In the 16,000-kc. band the daylight ground wave ranges were not in excess of 3 miles

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whereas the night ranges extended out to 22 miles in the same frequency band. These results can only be interpreted by supposing that there was some agency at work which produces a markedly greater absorption of the ground wave in the day-time than it does during the dark hours. If this were not true we would be forced to assume that we had to deal with some new kind of a sky wave produced by a very low refracting or reflecting layer. The absence of fading, however, makes the second possibility less likely. The writer is inclined to the opinion that the absorption of these very high frequencies must be greater in the day-time than during the dark hours even for those portions of the wave front which travel close to the earth and which are commonly spoken of as ground waves. No information is available from these particular experiments as to the state of polarization of these ground waves.

It might be that this information is only in line with what we know of somewhat longer wavelengths. A distance of 22 miles for a frequency of 16,000 kc. might be considered as comparable to the distance of 342 miles for a frequency of 1,000 kc. and of course it is well known that 1,000 kc. would suffer enormous variations in intensity at a distance of 342 miles, if one compared the night and day signals; but in the past it has been pretty generally assumed that at 1,000 kc. at a distance of 342 miles, the differences are due to the different character of the sky waves in the two cases.

The Austin-Cohen transmission formula is not to be applied for night reception of 1,000 kc. over the distances mentioned on account of the wave paths having been so utterly different in the two conditions of night and day. In other words, the sky wave, owing to the greater height of the refracting layer, is much more effective at night but in the case of the high frequency ground waves, which are the subject of this note, it is difficult to see how they can have had anything to do with sky waves at all unless they are waves of a variety hitherto unknown or at least unrecognized.

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A SUGGESTION OF A CONNECTION BETWEEN RADIO FADING AND SMALL FLUCTUATIONS IN THE EARTH'S MAGNETIC FIELD*

By
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The fading of radio signals has been referred to time and again as an interference effect. The suggestion that the upper atmosphere may be an important factor in enabling electromagnetic waves to propagate over appreciable distances has been the origin and support of this view. The wave traveling along the ground and the wave arriving by reflection have been supposed to interfere. Within the last few years, however, it has become obvious that more than simple interference is involved. Thus short-wave transmission, i. e., transmission on wavelengths below 100 meters in length, showed in the hands of Taylor and Hulburt[1] that beyond a certain range signals are received entirely by an overhead route. In spite of this fact, the signals show fading. Part of this is explained as a general shift in the height of the reflecting layer. However, this can be by no means all because much of the fading is rapid and does not correspond to the disappearance of the signal in the vicinity of the receiver such as occurs at the skip distance. In fact, the experiments by Mr. Tuve and the writer[1] on reflection of waves from the upper atmosphere showed that fading exists for the reflected waves quite independently of interference with the ground wave. Two suggestions have been made at the time as to a possible cause of the fading. One of them supposed the down-coming wave to be a result of the interference of two or more waves. During a discussion of the subject held at the Department of Terrestrial Magnetism, Dr. A. H. Taylor suggested that the Earth's magnetic field might be the cause of some of the

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[1]For references see bibliography.

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fading. On examining the matter, the writer found t
since the upper atmosphere must show magnetic dou
refraction and since, therefore, there must be in gen
at least two possible rays between the transmitter and
ceiver, a change in the interference conditions of these
rays must take place whenever the Earth's magnetic
changes. The change in the relative phase of the two
turns out to be large even if the Earth's field ch
slightly. Several gammas (0.00001 Gauss) may tu
plane of polarization of the ray through 30 deg. As a
the vertical component of the electric intensity n
changed from 0 to its maximum value in a field
gammas. The interference effects here discussed ar
the passage of the ray through an appreciable thic
ionized air. If the rays are reflected, the interferen
can not be nearly as pronounced. A study of the m
of fading in connection with small fluctuation
Earth's field may lead, therefore, to an underst
the manner in which the ionization increases
height.

Some simple calculations are given below i
show how the interference conditions may chan
same time the effect of a change of frequency
ionization are estimated. The calculations do n
to be exact. They are intended to show the orde
tude of the effects.

Before performing the calculations we shal
assumptions about the nature of propagation of
which will be used. We suppose the upper a
be partly ionized into electrons as well as hea
neglect the effect of heavy ions and consider
electrons alone. The density of free electro
to increase with the height up to a certain le
netic field of the Earth affects the motion of
manner discussed by Appleton[3] as well a
Schelleng[4]. We calculate below the effects o
of received signals due to ch in the i
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leading to a solution of the type

$$y = \frac{\cos\theta_0}{a} \sin \frac{ax}{\sin\theta_0} \tag{5}$$

The phase of the wave at any point is $\varphi = \int \frac{2\pi\nu}{V} ds$, the integral being taken over the path followed by the ray. If φ changes by π, the interference of the ray with another one of constant phase is changed from constructive to destructive. We are interested, therefore, in the phase difference $\Delta\Phi$ for the two rays if we are to determine the polarization of the wave at the receiving station. A change in the Earth's field δH produces a change $\delta\Delta\varphi$ in the phase difference $\Delta\varphi$. If this change $\delta\Delta\varphi = \frac{\pi}{2}$, the direction of polarization of the resultant linear vibration is turned through $90°$. We must obtain φ for either ray on the supposition that the position of the receiving and the transmitting stations is kept fixed. If the distance between them be l, formula (5) determines $\theta_0 = \frac{al}{\pi}$. Now, in general, if μ is the refractive index

$$\varphi = \frac{2\pi\nu}{c}\int \mu ds = \frac{2\pi\nu}{c}\int \sin e_0 \left(1 + \left(\frac{dy}{dx}\right)^2\right) dx \tag{6}$$

provided the refractive index varies only with the height and provided it is unity at the surface of the ground. Using (5)

$$\varphi = \frac{2\pi\nu}{c}\, l\, \frac{1+\sin^2\theta_0}{2\sin\theta_0} = 2\pi\, \frac{1+\frac{a^2l^2}{\pi^2}}{2\left(\frac{al}{\pi}\right)}\, \frac{l}{\lambda} \tag{7}$$

$$\left(\lambda = \frac{c}{\nu} = \text{wavelength}\right)$$

Keeping the range l and the frequency ν fixed, we suppose the Earth's field H to have undergone a variation δH. This changes a^2 on account of a change in ν_0 (see (4)). The change in φ due to δa is

$$\delta\varphi = \pi\, \frac{l}{\lambda}\, \frac{\frac{a^2l^2}{\pi^2} - 1}{\frac{al}{\pi}}\, \frac{\delta a}{a} = \pi\, \frac{l}{\lambda}\, \frac{A^2l^2 - 1}{Al}\, \frac{\delta A}{A} \tag{8}$$

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where $\pi A = a$, and

$$\delta\,\Delta\varphi = \delta\,(\varphi_2 - \varphi_1) = \pi\,\frac{l}{\lambda}\left(\frac{A_2 l^2 - 1}{A_2 l}\,\frac{\delta A_2}{A_2} - \frac{A_1 l^2 - 1}{A_1 l}\,\frac{\delta A_1}{A_1}\right)$$

It follows from (4) that $2\,\frac{\delta a_1}{a_1} = -\frac{v_0}{v + v_0}\,\frac{\delta v_0}{v_0}$, $2\,\frac{\delta a_2}{a_2} = \frac{v_0}{v - v_0}\,\frac{\delta_0}{v_0}$. Remembering also that $H^{-}\,\delta H = v_0^{-1}$ we have

$$\delta\,\Delta\varphi = \frac{\pi\,l}{2\,\lambda}\left(\frac{A_2^2 l^2 - 1}{A_2 l}\,\frac{v_0}{v - v_0} + \frac{A_1^2 l^2 - 1}{A_1 l}\,\frac{v_0}{v + v_0}\right)\frac{\delta H}{H} \tag{9}$$

Let $l = 0$, then

$$\delta\,\Delta\varphi = -\frac{\pi}{2}\,\frac{1}{\lambda}\left(\frac{1}{A_2}\,\frac{v_0}{v - v_0} + \frac{1}{A_1}\,\frac{v_0}{v + v_0}\right)\frac{\delta H}{H}$$

The quantity A_2 may be obtained in this case from the height h to which the wave goes. It is easily shown to be for this case $A_2 = \frac{1}{\pi h}$ so that for zenith reflection

$$\delta\,(\Delta\varphi) = -\frac{\pi}{2}\cdot\frac{\pi h}{2}\,\frac{v_0}{v - v_0}\left(1 + \left(\frac{v - v_0}{v + v_0}\right)^{1/2}\right)\frac{\delta H}{H} \tag{10}$$

If $h = 100$ kilometers and $\lambda = 70$ meters, then $\frac{v_0}{v - v_0} \tilde{=} 1/2$ and $\delta\,(\Delta\varphi) \tilde{=} -3.8 \times 10^{3}\,\frac{\pi}{2}\,\frac{\delta H}{H}$. Hence a change in H by $\frac{1}{3800}$ part of its whole amount would turn the plane of polarization of the downcoming beam through 90 deg. This would cause very severe fading even at a short distance from the transmitter for a change of 13γ ($= 13 \times 10^{-5}$ *Gauss*) in the Earth's field. A field one-third of that amount would cause appreciable fading. Variations of that order of magnitude in the Earth's field are very frequent and may, therefore, cause an appreciable amount of fading.

Although formula (10) applies strictly to the case of zenith reflection, it is a good approximation to (9) for considerable ranges because $A_2 l$ becomes equal to unity [for the number chosen above] if $l = 300$ kilometers. For 150 kilometers (10) is still a reasonable approximation. This suggests a connection with the distance of maximum fading for broadcast wavelengths. This may be determined in addi-

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tion to the proportion of ground and reflected waves by the fact that for a certain l, $\delta\Delta\varphi$ in accordance with (9) must vanish. Since in formula (10) the phase velocities (2), (3) have been used and since these velocities cannot be applied well for short-range transmission, we should expect the fading to be greatest some place between $l=0$ and $l=300$ kilometers.

Formula (7) gives also the change in the condition for interference if the magnetic field of the Earth stays constant and the frequency of the signal is changed. Using a' as given in (4) we get $2\,\frac{\delta a_2}{a_2}=-\frac{2v-v_0}{v-v_0}\,\frac{\delta v}{v}$,

$2\,\frac{\delta a_1}{a_1}=-\frac{2v+v_0}{v+v_0}\,\frac{\delta v}{v}$ and hence

$$\left.\begin{aligned}
\delta\varphi_2&=\frac{\pi l}{\lambda}\left(\frac{2}{A_2 l}-\frac{A_2^2 l^2-1}{2A_2 l}\,\frac{v_0}{v-v_0}\right)\frac{\delta v}{v}\\
\delta\varphi_1&=\frac{\pi l}{\lambda}\left(\frac{2}{A_1 l}+\frac{A_1^2 l^2-1}{2A_1 l}\,\frac{v_0}{v+v_0}\right)\frac{\delta v}{v}\\
\delta(\varphi_1-\varphi_1)&=\frac{2\pi l}{\lambda}\left(\frac{1}{A_2 l}-\frac{1}{A_1 l}+\frac{1-A_2^2 l^2}{4A_2 l}\,\frac{v_0}{v-v_0}+\right.\\
&\left.\frac{1-A_1^2 l^2}{4A_1 l}\,\frac{v_0}{v+v_0}\right)\frac{\delta v}{v}
\end{aligned}\right\}\quad(11)$$

For short range transmission this becomes

$$\delta(\varphi_2-\varphi_1)=\frac{\pi}{2}\,\frac{\pi h}{\lambda}\left[-4\left(\sqrt{\frac{v+v_0}{v-v_0}}-1\right)+\frac{v_0}{v-v_0}\left(1+\sqrt{\frac{v-v_0}{v+v_0}}\right)\right]\frac{\delta v}{v}\qquad(12)$$

With $\lambda=70$ meters and $h=100$ kilometers $\delta(\varphi_2-\varphi_1)=\frac{\pi}{2}\times 1400\,\frac{\delta v}{v}$. Hence if $\frac{\delta v}{v}=\frac{1}{1400}$ the direction of polarization is rotated through 90°. This corresponds to a frequency change of 3000 cycles per second. Hence even a frequency change of 1000 should produce an appreciable effect on the intensity of the received signal.

Finally, we estimate for the above case the change in $\varphi_2-\varphi_1$ which is produced by a change of ionization. We find in a way quite similar to that just used that

$$\delta(\varphi_2-\varphi_1)=\frac{\pi}{2}\,\frac{l}{\lambda}\left(\frac{A_2^2 l^2-1}{A_2 l}-\frac{A_1^2 l^2-1}{A_1 l}\right)\frac{\delta N}{N}\qquad(13)$$

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where δN is the change in the number of electrons per unit volume. For short-range transmission this becomes

$$\delta(\varphi_2 - \varphi_1) = -\frac{\pi}{2}\frac{\pi h}{\lambda}\left(-1 + \sqrt{\frac{v+v_0}{v-v_0}}\right)\frac{\delta N}{N} \qquad (14)$$

With the numbers used above, a rotation through 90° is produced if $\frac{\delta N}{N} = \frac{1}{1800}$. Since changes in the ionization are presumably responsible for changes in the Earth's field, it is reasonable to suppose that (9) and (14) should be combined in any actual case in order to give the resultant change.

In the above we have constantly referred to the short wavelength region ($v > v_0$). If the wavelength is long, i.e., if $v < v_0$, only the ray defined by (3) returns to the Earth. If this ray interferes with the ground, wave effects quite similar to those just discussed are produced. Formulas (9), (11), (13) must be modified, however, by omitting the terms in A_2. The order of magnitude of the effects is still the same.

Even though $\delta(\varphi_2 - \varphi_1)$ is the same for a considerable range according to (9), (11), (13), the fading produced may vary considerably over that range. Thus the value of $\varphi_2 - \varphi_1$ is very important because it determines whether the electric vector of the downcoming wave is horizontal or not. The same fluctuation of H may produce, therefore, a reenforcement of the signal in one place and a diminution in another.

Small frequency changes are, of course, known to produce an effect on the intensity of the received signal. It is also known that interrupted continuous waves show less fading on short wavelengths than continuous waves[5]. It is somewhat difficult to point out definitely what has been happening in the experiments just mentioned. For short wavelengths, however, it seems certain that the fading must be entirely due to the condition of the wave returned by the upper atmosphere. The high "resolving power" of the interference apparatus in that case demonstrates a behavior similar to the one discussed above. It cannot be interpreted on the basis of interference with the ground wave. It must be also remembered that experiments by Mr. Tuve and the writer (loc. cit.) showed that the reflected wave shows fad-

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ing which is sufficiently great to explain an appreciable amount of the general fading.

The influence of magnetic storms on transmission has given opportunity for considerable discussion. However, it would be dangerous to maintain that the simple calculations given above explain the effects observed. Presumably magnetic storms are accompanied by large changes in the ionization of the reflecting layer. A complete explanation must, therefore, be based on formulas corresponding to both (9) and (13).

SUMMARY: The state of polarization of a radio wave returned by the reflecting layer is studied as a function of (a) small changes in the intensity of the Earth's magnetic field, (b) small changes in the frequency of the wave, and (c) small changes in the ionization of the atmosphere.

It is shown that for the special electron distribution considered appreciable effects on the intensity of the signal are to be expected for fluctuations in the Earth's field of the order of 5 gammas, for changes in the frequency of the order of 1,000 cycles, and for changes in the ionization of the order of one part in 5,000. [The number applies to $\lambda = 70$ meters.]

The dependence of fading on range indicates that there is a certain range of maximum fading which is of the general order of 100 miles.

APPENDIX

It seems that a certain amount of justification is needed for using Snell's law and for discussing two rays only. This is obtained by the following considerations.

1. *Equations of propagation in a homogeneous magnetically active medium.* The equations of propagation in a magnetically active medium are generally well known. However, it seems that in their applications to radio some misconceptions have arisen. For this reason it will be well to summarize the important facts. It is supposed that the wave propagating through the layer is monochromatic and has a frequency $\nu = \frac{\omega}{2\pi}$. The medium through which it propagates is supposed to contain N electrons per cm^3. Each electron has a charge e and a mass m. The forces acting on

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each electron are as follows: (a) The electric intensity $(E_x, E_y, E_z)\, e^{iwt}$ of the wave, (b) the force $\frac{4\pi}{3}(P_x, P_y, P_z)$ where P is the polarization of the medium, (c) the force due to the Earth's magnetic intensity (O, O, H_o) supposed here to be constant and directed along the axis OZ, the magnitude of this force being $\frac{e}{c}(y, -x, O)H_o$, and (d) the forces which are brought about by collisions.

Denoting the displacement of an electron from its position of equilibrium by (x, y, z) we have

$$(P_x, P_y, P_z) = Ne(x, y, z)$$

Thus the equations of motion are neglecting (d).

$$\begin{aligned} x &= \frac{e}{m}E_x e^{iwt} + \frac{e}{mc}yH_o + \frac{4\pi}{3}\frac{Ne^2}{m}x \\ y &= \frac{e}{m}E_y e^{iwt} - \frac{e}{mc}xH_o + \frac{4\pi}{3}\frac{Ne^2}{m}y \qquad (1) \\ z &= \frac{e}{m}E_z e^{iwt} + \frac{4\pi}{3}\frac{Ne^2}{m}z \end{aligned}$$

For most purposes we can neglect the terms in $\frac{4\pi}{3}\frac{Ne^2}{m}$ with the result that Maxwell's equations become

$$\begin{aligned} \frac{\delta H_z}{\delta y} - \frac{\delta H_y}{\delta z} &= \frac{1}{c}\left[(1+\alpha)\frac{\delta E_x}{\delta t} - i\beta\frac{\delta E_y}{\delta t}\right] \\ \frac{\delta H_x}{\delta z} - \frac{\delta H_z}{\delta x} &= \frac{1}{c}\left[i\beta\frac{\delta E_x}{\delta t} + (1+\alpha)\frac{\delta E_y}{\delta t}\right] \qquad (2) \\ \frac{\delta H_y}{\delta x} - \frac{\delta H_x}{\delta y} &= \frac{1+\gamma}{c}\frac{\delta E_z}{\delta t} \end{aligned}$$

where $\alpha = \frac{4\pi Ne^2}{m(\omega'^2 - \omega^2)}$, $\beta = \frac{4\pi Ne^2}{m}\frac{\omega'}{\omega(\omega'^2 - \omega^2)}$,

$\gamma = -\frac{4\pi Ne^2}{m\omega^2}$, and $-\frac{1}{c}\frac{\delta H}{\delta t} = curl\, E$.

If a plane wave is possible, all the quantities involve $exp\, \frac{2\pi i}{\lambda}(lx + my + nz - Vt)$. Substituting in Maxwell's equations we have:

$$\frac{V}{c}H_x = mE_z - nE_y;\quad mH_z - nH_y = -\frac{V}{c}[(1+\alpha)E_x - i\beta E_y]$$

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$$\frac{V}{c}H_y = nE_x - lE_z;\ nH_x - lH_z = -\frac{V}{c}[i\beta E_x + (1+\alpha)E_y] \tag{3}$$

$$\frac{V}{c}H_z = lE_y - mE_x;\ lH_y - mH_x = -\frac{V}{c}(1+\gamma)E_z$$

whence

$$\frac{E_x}{n(1A - imB)} = \frac{E_y}{n(mA + ilB)} = \frac{E_z}{A^2 - B^2 - (l^2+m^2)A}$$

$$= \frac{\frac{V}{c}H_x}{m(A^2 - B^2 - A) - in^2 lB} = \frac{\frac{V}{c}H_y}{l(B^2 + A - A^2) - in^2 mB}$$

$$= \frac{\frac{V}{c}H_z}{in(l^2+m^2)B} \tag{4}$$

where

$$A = 1 - \frac{V^2}{c^2}(1+\alpha);\ B = \beta\frac{V^2}{c^2};\ C = 1 - \frac{V^2}{c^2}(1+\Upsilon)$$

These equations determine the polarization of the wave if l, m, n, V are known. In order to find V, eliminate E_x, E_y E_z, H_x, H_y, H_z in (3). Then we find

$$(A^2 - B^2)(C - n^2) + (n^2 - 1)AC = 0 \tag{5}$$

or letting

$$u = \frac{V^2}{c^2} \tag{6}$$

we have

$$u(pu^2 + qu + r) = 0 \tag{7}$$

where

$p = (1+\gamma)[(1+\alpha)^2 - \beta^2 = (1+\gamma)[1 + \alpha + \alpha(1+\gamma)]$

$q = -2(1+\alpha)(1+\gamma) + (n^2-1)[(1+\alpha)(\alpha-\gamma) - \beta^2] = -2(1+\alpha)(1+\gamma) + (n^2-1)(\alpha-\gamma)$

$r = 1 + \gamma + (n^2-1)(\gamma-\alpha) = 1 + \gamma + (n^2-1)(\gamma-\alpha)$

For every n there are two roots of (7) which are not identically zero. These roots correspond to two possible velocities of propagation in a direction making a given angle $\cos^{-1}n$ with the field H. The roots are real if $q^2 > 4pr$ and complex if $q^2 < 4pr$. Thus either both waves are possible or none.

2. *Propagation of waves in a non-homogeneous medium.* We shall suppose next that the concentration of electrons

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varies from point to point in the medium in a known manner. Thus N is a known function of x, y, z. We shall further suppose that within a wavelength $N(x, y, z)$ is practically constant. Under these conditions we can discuss the propagation of waves in the medium by means of geometrical optics to a certain extent.

Let us suppose first for simplicity that the magnetic field is absent so that the medium is isotropic but non-homogeneous. This case is well known and forms the basis of Hamilton's theory of dynamics. The fundamental equation to solve is the wave equation

$$\frac{\delta^2 u}{\delta x^2} + \frac{\delta^2 u}{\delta y^2} + \frac{\delta^2 u}{\delta z^2} - \frac{1}{[V(x, y, z)]^2} \frac{\delta^2 u}{\delta t^2} = 0 \qquad (8)$$

subject to the condition that u is a periodic function of the time having a frequency $\nu = \frac{\omega}{2\pi}$. This reduces (8) to

$$\Delta u + \frac{\omega^2}{V^2} u = 0 \qquad (8')$$

The statement that geometrical optics give an approximate solution of the problem is equivalent to saying that

$$u = C \exp 2\pi i\nu \left(t - \int \frac{ds}{V}\right) \qquad (9)$$

is a sufficiently exact solution of (8') where $\int \frac{ds}{V}$ is the path difference expressed in seconds, the integral being taken over a possible ray, i. e., subject to the condition

$$\delta \int \frac{ds}{V} = 0 \qquad (10)$$

by Fermat's Principle. Substituting (9) into (8) and letting

$$S = \int \frac{dx}{V} \qquad (11)$$

we have

$$\frac{i}{2\pi\nu} \Delta S + \left[\left(\frac{\delta S}{\delta x}\right)^2 + \left(\frac{\delta S}{\delta y}\right)^2 + \left(\frac{\delta S}{\delta z}\right)^2\right] - \frac{1}{V^2} = 0 \qquad (12)$$

Fermat's Principle and the construction of Huygens being equivalent, it is found without difficulty that

$$\left(\frac{\delta S}{\delta x}\right)^2 + \left(\frac{\delta S}{\delta y}\right)^2 + \left(\frac{\delta S}{\delta z}\right)^2 - \frac{1}{V^2} = 0 \qquad (13)$$

so that (12) is satisfied provided ΔS may be neglected in

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comparison with $\frac{2\pi v}{V^2}$ or provided the change in $\frac{1}{V}$ per unit length when multiplied by $\frac{V^2}{2\pi V}$ is negligible compared to unity [because $\frac{\delta^2 S}{\delta x^2}$ is of the order of $\frac{\delta}{\delta x}\left(\frac{1}{V}\right)$]. This means that $\left|\frac{1}{2\pi v}\frac{\delta V}{\delta x}\right| << 1$ or $\left|\frac{\lambda}{2\pi}\frac{\delta \log V}{\delta x}\right| << 1$. Thus the relative change in V in $\frac{1}{2\pi}$ fraction of a wavelength must be negligible in comparison with unity.

The above reasoning is well known and has been made use of recently by Louis de Broglie[4] in a quite different connection. So far it applies only to non-polarized waves. Our first problem is to see what happens in the case of polarized electromagnetic waves. The equations of propagation are

$$curl\ H = \frac{\epsilon(x,y,z)}{c}\frac{\delta E}{\delta t} \tag{14}$$

$$curl\ E = -\frac{1}{c}\frac{\delta H}{\delta t} \tag{15}$$

where ϵ is the effective dielectric constant of the medium, viz, $1 + 4\pi\frac{P}{E}$. This ϵ is supposed to vary from point to point. However, the variation in ϵ is supposed to be slow so that $\frac{\delta \log \epsilon}{\delta x} << \frac{\delta Ex}{\delta x}$. Since the components of E and H satisfy the wave equation with $V^2 = \frac{c^2}{\epsilon}$ we could try to use (9) for the components directly. However, we must satisfy not only (14) and (15) but also the relations

$$div\ H = C \tag{16}$$

$$div\ D = div\ (E + 4\pi P) = div\ (\epsilon E) = 0 \tag{17}$$

as well as (14) and (15). Solutions given by (9) will not necessarily satisfy either of these equations. It is known, however, that a scalar φ and a vector a each of which satisfies (9) can be used to express E and H as

$$\vec{E} = -\frac{1}{c}\frac{\delta \vec{a}}{\delta t} - \Delta\varphi \tag{18}$$

$$\vec{H} = curl\ \vec{a} \tag{19}$$

provided

$$\Delta\varphi + \frac{1}{c} div\frac{\delta \vec{a}}{\delta t} = 0 \tag{20}$$

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and provided ϵ satisfies the condition of varying slowly [as defined above]. The proof of this is quite similar to that in the well-known case of constant ϵ and need not be given. An equation equivalent to (20) is

$$\frac{c}{V^2}\frac{\delta\varphi}{\delta t} + div\,\vec{a} = 0 \qquad (20')$$

which is well known for $V = c$.

The case of a plane polarized wave in a homogeneous medium can be obtained by letting

$$a_x = -\frac{cE_0}{\omega} sin\,\omega\left(t - \frac{z}{c}\right),\ a_y = a_z = 0;\ \varphi = 0 \qquad (21)$$

This wave travels along OZ, and has its electric vector along OX, while the magnetic vector is along OZ.

A plane wave making an angle θ with 03 and polarized with its electric vector along OX is similarly represented by

$$a_x = -\frac{cE_0}{\omega} sin\,\omega\left(t - \frac{z\cos\theta + y\sin\theta}{c}\right),\ a_y = a_z = 0;\ \varphi = 0 \qquad (22)$$

The wave normal for this wave is the line $\frac{z}{\cos\theta} = \frac{y}{\sin\theta}$. The solution (22) applies to a homogeneous medium only. If the medium is not homogeneous, we apply solution (9) to a_x so as to satisfy (8). Equations (14), (15), (16), (17) are then also satisfied if (18), (19) are used. Thus this wave travels without change of polarization. The amplitude of the electric intensity is found to remain constant along the path, while the magnetic intensity at any point is $\frac{c}{V}$ of the original. This corresponds also to the fact that the ratio of the electric and magnetic intensities in the wave is $\frac{V}{c}$.

Suppose next that the wave normal is in the xz plane. For a homogeneous medium

$$\begin{aligned} a_x &= -\frac{c\,E_0\cos\theta}{\omega} sin\,\omega\left(t - \frac{z\cos\theta + x\sin\theta}{c}\right) \\ a_y &= 0 \\ a_z &= +\frac{c\,E_0\sin\theta}{\omega} sin\,\omega\left(t - \frac{z\cos\theta + x\sin\theta}{c}\right) \end{aligned} \qquad (23)$$

In order to find the solution for the non-homogeneous medium [the non-homogeneity being confined to the direction z]

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we suppose s to be determined by (10) and the quantity $-\frac{c E_0}{\omega} \sin \omega \left(t - \frac{z \cos \theta + x \sin \theta}{c} \right)$ to be the value of u for $V = c$. We then try

$$a_x = u \cos (sz), \; a_y = 0, \; a_z = -u \sin (sz) \qquad (24)$$

$$u = -\frac{c E_0}{\omega} \sin \omega \left(t - \int \frac{ds}{V} \right)$$

Introducing (24) into (8) we are concerned with expressions of the type $\frac{\delta^2 u}{\delta x^2} \cos (sz) + 2 \frac{\delta u}{\delta x} \frac{\delta}{\delta x} (\cos (sz)) + u \frac{\delta^2}{\delta x^2} \cos (sz)$. The particular non-homogeneity assumed makes $\frac{\delta}{\delta x} \cos (sz) = 0$. However $\frac{\delta}{\delta z} (\cos sz) \neq 0$. This contributes a term of the type $\frac{\delta u}{\delta z} \frac{\delta (\cos sz)}{\delta z}$. But $\cos (sz) = V \frac{\delta S}{\delta z}$ and $\frac{\delta}{\delta z} \cos (sz) = \frac{\delta V}{\delta z} \frac{\delta S}{\delta z} + V \frac{\delta^2 S}{\delta z^2}$. So we have to consider in (12) additional terms $\frac{i}{2\pi\nu} \left[\frac{\delta V}{\delta Z} \left(\frac{\delta S}{\delta z} \right)^2 + V \frac{\delta S}{\delta z} \frac{\delta^2 S}{\delta z^2} \right]$. A consideration quite similar to the one which made us disregard ΔS makes us disregard these terms as well.

The propagation of a wave in an isotropic medium thus presents no difficulty. If the medium is rendered non-isotropic through the influence of a magnetic field and if the terms in $\frac{4\pi}{3} \frac{Ne^2}{m}$ occurring in (1) may be neglected, the two possible velocities of propagation in a given direction (l, m, n) may be determined by means of (7) and the states of polarization corresponding to each of these velocities are obtained by (4). If it is at all possible to discuss the propagation of waves by means of rays, then the rays must obey Fermat's Principle of least time. Therefore, we arrive at the following synthetic solution of the problem. While the plane wave is in a region of the atmosphere in which the electron density is zero, we may consider it resolved into two plane waves with such polarizations that on entering the refracting portions of the atmosphere each of these has a

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polarization satisfied by (4). From there on each wave is supposed to be bent in accordance with Fermat's Principle keeping its polarization determined by the particular root of (7) chosen for it in the beginning.

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NOTE ON PIEZOELECTRIC GENERATORS WITH SMALL BACK ACTION

By
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A number of circuit arrangements suitable for automatic piezo control have been devised. Three such arrangements are shown in the Figures.

In Fig. 1, the piezoelectric element is connected between the grid and a copper cylinder which is free to move over

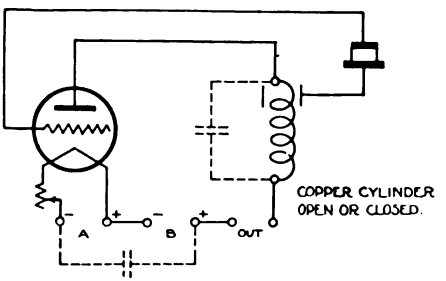

Figure 1

the coil in the plate circuit. The copper cylinder can be closed (forming a single turn) or open (axial cut.) The latter is more efficient since no appreciable power dissipation can take place in the cylinder, while the first method provides good tuning over a considerable frequency range. Experiments have shown this arrangement to be useful. The back action is very small since the piezoelectric element is coupled through distributed capacity to the plate branch.

In the arrangement of Fig. 2, two tubes are used. Tube 1 contains the ordinary piezo oscillator. Tube 2 delivers the

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load. The same grid variations take place as on the grid of tube 1 but without loading it appreciably.

In Fig. 3, the same scheme is utilized with the exception

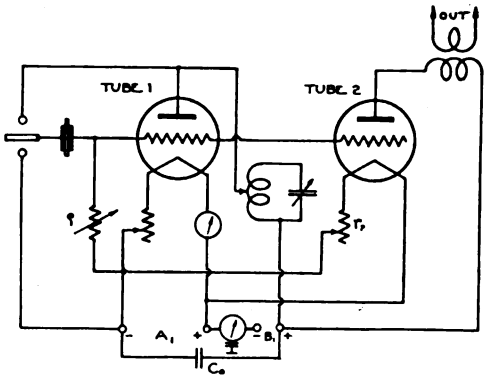

Figure 2

that a coupling condenser C is used between the two grids and a grid path r and a voltage in series are inserted. Additional plate voltage taken from B_2 can be used for in-

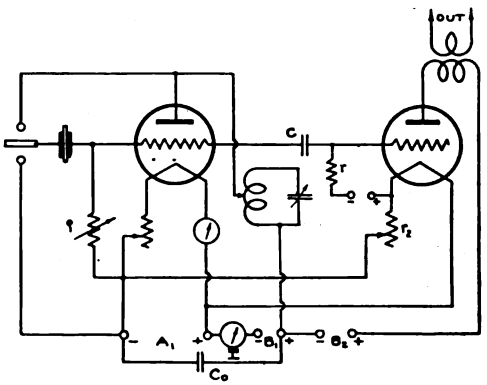

Figure 3

creasing the output. With this circuit the last tube can be adjusted to the proper point of the dynamic characteristic while the leak ρ of the first tube is used for producing proper amplitude of the current in that circuit.

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STANDARD FREQUENCY DISSEMINATION*

By
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In the radio laboratory of the Bureau of Standards, the term "standard frequency dissemination" designates that phase of the work devoted to making frequency standards widely available. The Bureau does this in a number of ways including the calibration of piezo oscillators and frequency meters, but this paper deals with the standard frequency disseminations through the medium of radio transmission.

The basis of this scheme depends upon the fact that a standard of radio frequency may, neglecting the effects of interference, be transmitted over great distances and reproduced at the receiving station with an accuracy equal to that attainable at the source of the frequency. Hence, if a radio transmitting station is operating on a constant frequency of accurately known value, that station serves as a disseminator of the frequency over an area determined by the effective range of transmission and reception.

The Bureau of Standards has three avenues for this means of standard frequency dissemination. First, standard frequency transmissions; second, selection by actual frequency measurements of certain transmitting stations which are termed "standard frequency stations;" third, the selection of certain "constant frequency stations" which maintain their frequencies close to the licensed values.

The standard frequency signals are transmitted from the Bureau's station WWV and from other stations which are equipped to make the transmissions with accuracy and regularity and which employ frequency standards in agreement with those used at the Bureau of Standards. Announcements of these transmissions are made from time to time in the Radio Service Bulletin, in the newspapers and in

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radio magazines. The transmissions from the Bureau of Standards have been maintained approximately once each month since 1922. Prior to October, 1924, the range of frequencies covered was from 125 to 2,000 kilocycles, but since that date this has been extended to 6,000 kilocycles. Evidence of the importance and general use of these transmissions was given by the fact that an announcement of their possible discontinuance in April, 1926, brought many objections from various laboratories.

In cooperation with the transmissions from WWV, similar transmissions were established from station 6XBM at Stanford University, California, in September, 1924. A standard frequency meter was built at the Bureau of Standards, given an initial calibration, and shipped to the University. A final calibration in terms of the Bureau's frequency standard was obtained by means of harmonics based upon a few fundamental frequency values. The values were determined by simultaneous measurements of transmitting stations by observers at the Bureau and Stanford University and from a small frequency meter having a number of fixed points determined in the Bureau's laboratory. The transmissions from 6XBM were terminated in June, 1926.

During the past year and in cooperation with the Bureau of Standards, standard frequency transmissions have been established from station 1XM, Massachusetts Institute of Technology, and from station 9XL, Gold Medal Flour Co., near Minneapolis, Minn. These transmissions cover the higher frequencies of direct value to amateurs.

Recently the accuracy of the standard frequency transmissions from WWV has been somewhat increased. Heretofore the transmission frequencies were determined by reference to primary standard frequency meters having high precision and constancy in calibration. The transmissions are now based upon a single frequency value of a quartz plate used in a piezo oscillator, checked during transmission by a standard frequency meter. A special advantage of the piezo oscillator as compared with the frequency meter is that it is unaffected by any power fluctuations which may occur in the transmitting set, but a change in the transmitted frequency is manifested in the head phones connected to the piezo oscillator by a deviation from zero beat.

The measurements upon standard frequency stations

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have been regularly maintained since November, 1923, and the results of these measurements are published monthly in the Department of Commerce Radio Service Bulletin. Only stations sufficiently close to the Bureau of Standards laboratory to be regularly and reliably received are measured. The number of stations that can be listed is necessarily limited. At the present time there are thirteen standard frequency stations, and these lie within the frequency range from 17 to 1,000 kilocycles.

For the measurements of the frequencies of the standard frequency stations the Bureau of Standards now has two completely equipped laboratories. One of these is located at the Bureau, while the other is situated near Kensington, Maryland, in a location which is almost entirely free from interference from electrical machinery and electrical apparatus.

The apparatus for measuring the frequencies of distant transmitting stations comprises a receiving set, a local generator, and a frequency meter. The generator is tuned to zero beat, and it is then exactly reproducing the frequency of the distant station subject to a slight error due to an audible beat in the band of audio frequencies approximately 32 cycles in width. The frequency of the local generator is then measured by means of the frequency meter. In the case of measurements of transmitting stations having frequencies of approximately 25 kilocycles or less, the width of the audio frequency band in the region of zero beat is sufficiently wide to introduce objectionable error in the measurements. To overcome this, a method is employed which involves adjusting the variable condenser of the local generator so that an audible beat of equal pitch is heard each side of the condenser setting corresponding to exact zero beat. A type of variable condenser is employed such that the changes in dial setting are directly proportional to the change in its capacity. The true zero-beat setting of the condenser is therefore located half-way between the settings giving the beat of equal pitch. Another method is to adjust the local generator until the beat produced matches the beat from an audio tuning fork having a frequency of accurately known value.

Improvements in the accuracy with which station frequencies can be measured have recently been made. One of

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these improvements is merely a refinement of the original method and involves the use of a frequency meter so constructed as to give high precision. A method of securing increased accuracy in the measurements which is somewhat new in its application is through the use of a heterodyne frequency meter in conjunction with a piezo oscillator. A heterodyne frequency meter is a generator constructed to give high precision and provided with calibration curves plotted to a large scale. In making a station measurement, the zero-beat adjustment is obtained in the same manner as before. The setting of the condenser of the heterodyne meter is then noted and although this is read to a high degree of precision, the frequency corresponding to it is not highly accurate. This is due to the fact that the calibration of the heterodyne meter does not remain constant. A correction for this frequency is obtained by utilizing harmonics from the piezo oscillator and the heterodyne meter. These points are then located upon the curve sheet and determine a new curve of high accuracy which is exactly similar in shape to the corresponding portion of the original curve. The intersection of this new curve with the station setting of the heterodyne meter is the true frequency of the station. The accuracy of this method may be extended to a very high order, the principal limitations being the accuracy of calibration of the piezo oscillator and the effects of temperature and humidity upon this calibration and the scale chosen for the calibration curves of the heterodyne meter.

Since the number of stations upon which frequency measurements can be made is limited, an additional scheme for compiling a list of stations that may serve as standards was adopted by the Bureau. These are the constant frequency stations listed each month in the Radio Service Bulletin. In April the list included approximately 5 per cent of the broadcasting stations in the United States. The selection of these stations depends upon the nature of the transmitting equipment which must not be subject to sudden changes in frequency, upon care and diligence on the part of the station operators, and the use of a special device for determining the station frequency. The special device may be a frequency indicator or a piezo oscillator.

The standard frequency dissemination work herein described is furthered by the issuance of mimeographed publi-

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cations on methods of utilizing station frequencies, and apparatus for station frequency regulation. The Letter Circulars, obtainable by request from the Bureau of Standards, are named below.

Letter Circular 171, "Methods and apparatus for measurement of the frequencies of distant radio transmitting stations."

Letter Circular 180, "Specifications for frequency indicator, Bureau of Standards Type B, for use in radio transmitting stations."

Letter Circular 186, "Specifications for portable piezo oscillator, Bureau of Standards Type N."

Letter Circular 214, "Requirements of constant frequency stations."

Letter Circular 223, "Use of the piezo oscillator in radio broadcasting stations."

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FORMULAS FOR THE CALCULATION OF THE CAPACITY OF ANTENNAS*

By
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(Union College)

A knowledge of the capacity of an antenna is of importance in calculating the natural wavelength of an antenna, for purposes of design and in the comparison of different types of antenna. Only the electrostatic capacity will be here considered. If the wavelength used is large, compared with the length of the antenna, this may be regarded as giving the effective capacity of the antenna.

The capacity of a conductor is defined as the quotient of the quantity of its charge by its potential. The potential is the sum of the potentials given it by its own charge and by the charges on the earth and on all the other conductors of the system. The effect of the charges on the earth is taken into account by including in the system image wires, one for each of the actual conductors. Each image wire is supposed to be placed as far below the surface of the earth as the conductor to which it corresponds is above the surface of the earth, and to carry a charge equal and of opposite sign to that upon the conductor, to which it corresponds.

Exact formulas for the capacity of a conductor are known in only a few instances, and unfortunately even the simplest case of a single horizontal cylindrical wire antenna would seem to offer great mathematical difficulties.

The known exact formula for the capacity of an isolated ellipsoid has sometimes been used, the longest dimension of the ellipsoid being taken equal to the length of the wire and the other two axes each equal to the radius of its cross section. However, the surface of such an ellipsoid does not coincide at all closely with that of the cylindrical wire, as the ends are approached, and in addition no account is taken of the effect of the earth.

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The formula for the capacity of two parallel wires of infinite length is well known, and thus, by considering one of these to represent the image wire, the solution is at hand for a horizontal wire whose length is very great compared with its height above the earth. This condition is, however, not satisfied with the usual single horizontal wire antenna, and thus the assumption that the capacity per unit length is the same as for an infinite wire of the same diameter, placed at the same height above the earth, gives only an approximate value for the capacity.

The charge distribution on an infinite horizontal wire is uniform in the direction of the axis. That is, the charge on the cylindrical surface intercepted between planes perpendicular to the axis and taken one cm. apart, is everywhere the same. Furthermore, the potential at an external point, due to the charge on the wire is very closely the same as though the same charge were distributed with a uniform linear density along the axis of the wire. In the case of a finite horizontal wire, on the other hand, the charge density is manifestly greater at the ends than at the center.

Supposing, however, that a charge be distributed uniformly over its cylindrical surface, the resulting potential may be calculated for different points of its surface. These potential values differ very little except for points quite close to the ends, and only at the extreme ends does the potential drop appreciably. The potential given to the wire by a uniform distribution on its image wire is practically constant over its whole length.

We may assume the value calculated for the middle point of the wire as fairly representative, and assume it to give an approximation to the true value which is better than that obtained from the formula for the infinite wire. This procedure was used in obtaining the formulas derived by the author and published in 1917.*

The variation along the wire of the potentials calculated on the assumption of a uniform charge density, shows that such a distribution could not be in equilibrium. The fall of potential toward the ends would cause charge to flow from the center toward the ends thus lowering the potential at the center, and raising it at the ends, until the whole wire would reach a uniform potential. This would have a value

*Bureau of Standards Circular 74, "Radio Instruments and Measurements"

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which would evidently be less than the value previously calculated for the center, and greater than that calculated for the ends.

We may assume, as was done by Howe,* that the average of the potentials calculated for a uniform charge density, taken over the conductor, is a good approximation to the true equilibrium potential. The following summary shows the differences between the various approximations above enumerated in the case of a horizontal wire 50 ft. long, 0.01 ft. in diameter, and placed 25 ft. above the earth.

	RELATIVE POTENTIALS	
	Isolated	Earth Effect Included
Infinite Wire Assumption	———	18.420
Isolated Ellipsoid	18.420	———
Uniform Charge Density, Center	18.420	17.458
Howe's Assumption	17.807	16.874

An idea of the accuracy of the Howe assumption has been obtained by the author of the present paper by a method of successive numerical approximations to the true distribution of charge and the equilibrium potential, in two special cases. For the wire above assumed, isolated in space, it appears that the Howe's approximation to the potential is 0.2 to 0.3 per cent too large. For the same wire, arranged in a vertical position, with its lower end only one foot from the ground, the error is about 1 per cent, and in the same direction as before.

Such calculations are laborious and time consuming, and each new combination of wires has to be treated as a special case. Thus, no general statement can be made as to the accuracy of the Howe approximation in the case of more complicated antennas. The author believes, however, that with the long thin wires usual in antennas, the Howe method of approximation gives an accuracy sufficient for engineering requirements.

Accordingly this method of solution has been used to derive the formulas of this paper. All the more common forms of antenna have been included. In those cases where a number of wires have been joined together, two methods may be followed. First, the Howe potential may be calculated for each wire, on the supposition that a uniform linear charge density has been given to the whole system. The

* Lond. Elec. 73, p. 829; 1914.

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equilibrium potential is taken as the mean of the Howe potentials of all the wires, weighting each value according to the length of the wire to which it corresponds. Or, as an alternative method, the various elements may each be supposed uniformly charged, but with a different value of charge density for each wire. The relation between these different charge densities is determined so as to give the same Howe potential to each wire. The two methods are in close agreement for long thin wires.

A collection of formulas found by the methods described, covering different antenna types, together with tables of constants to aid in the calculations, and tables of the capacity itself for certain simple antenna systems, has been issued as a letter circular by the Bureau of Standards.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued May 24, 1927---July 26, 1927

By

JOHN B. BRADY

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,629,685—RADIO TELEPHONY—W. T. DITCHAM, of Twickenham, England. Filed Dec. 18, 1920, issued May 24, 1927. Assigned to Radio Corporation of America.

1,629,727—ELECTRIC WAVE PRODUCING AND CHANGING DEVICE—H. J. MURRAY, of Brooklyn, N. Y. Filed April 1, 1921, issued May 24, 1927.

1,629,825—RADIO APPARATUS—J. J. W. KENAN and WILLIAM M. CADY, of Newark, New Jersey. Filed March 28, 1925, issued May 24, 1927.

1,629,826—RADIO APPARATUS—J. J. W. KENAN and WILLIAM M. CADY, of Newark, New Jersey. Filed March 28, 1925, issued May 24, 1927.

1,629,867—ELECTRICAL CONDENSER—J. F. HERMAN, of Chappaqua, New York. Filed March 26, 1925, issued May 24, 1927. Assigned to Straitline Radio Corp., of New York.

1,629,882—CONDENSER—W. H. PETIT, of Dayton, Ohio. Filed May 26, 1922, issued May 24, 1927. Assigned to Miami Radio Corp., Co.

1,630,227—APPARATUS FOR RECORDING ELECTRICAL SIGNALS—ALBERT H. TAYLOR, of Washington, D. C. Filed Jan. 2, 1924, issued May 24, 1927.

1,630,349—ARRANGEMENT FOR SUPPRESSING IDLE CURRENT IN RADIO TRANSMITTERS—A. MEISSNER, Berlin, Germany. Filed April 6, 1925, issued May 31, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie M. B. H.

1,630,354—RADIO SIGNALING APPARATUS—T. D. PARKIN, Chelmsford, England. Filed Dec. 2, 1921, issued May 31, 1927. Assigned to Radio Corp. of America.

1,630,383—DETECTOR—E. J. HAVERSTICK, Oakmont, Pa. Filed June 28, 1923, issued May 31, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,630,431—ELECTRON DISCHARGE DEVICE—W. G. HOUSEKEEPER, New York, N. Y. Filed July 29, 1922, issued May 31, 1927. Assigned to Western Electric Co.

1,630,443—MULTIPLE TRIODE VACUUM TUBE—F. S. McCULLOUGH, Wilkinsburg, Pa. Filed Mar. 15, 1923, issued May 31, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,630,859—CIRCUIT SYSTEM FOR RADIO FREQUENCY CURRENTS—R. S. MINER, West Hartford, Conn. Filed April 2, 1923, issued May 31, 1927. Assigned one-half to The C. D. Tuska Co.

1,630,885—VARIABLE CONDENSER—J. R. BARNHART, Lakewood, Ohio. Filed April 9, 1924, issued May 31, 1927. Assigned to Walter M. Scott.

1,630,980—RADIO INTERFERENCE ELIMINATOR— F. W. STEIN, Atchison, Kansas. Filed April 30, 1925, issued May 31, 1927.

1,630,021—THERMIONIC INDICATING MEANS RESPONSIVE TO LIGHT VARIATIONS—J. J. DOWLING, Bathgar, Dublin, Ireland. Filed Mar. 2, 1926, issued May 31, 1927.

1,631,035—RADIO TUBE—H. K. HUPPERT, San Francisco, Calif. Filed Dec. 21, 1925, issued May 31, 1927.

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1,631,100—VIBRATION DAMPING MEANS FOR VACUUM TUBES—C. C. LAURITSEN, St. Louis, Mo. Filed Feb. 25, 1926, issued May 31, 1927. Assigned to The Dyal Co.

1,631,226—VARIABLE CONDENSER—E. J. SCHRODER, Lowell, Indiana. Filed April 20, 1925, issued June 7, 1927.

1,631,360—VARIABLE CONDENSER—G. H. CLARK, Brooklyn, N. Y. Filed June 27, 1922, issued June 7, 1927. Assigned to Radio Corp. of America.

1,631,453—VARIABLE CONDENSER—R. L. ASPDEN, Chorley, England. Filed May 29, 1926, issued June 7, 1927.

1,631,599—VARIABLE CONDENSER—A. LEIB, Berlin, Germany. Filed Sept. 14, 1923, issued June 7, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie M. B. H.

1,631,636—VARIABLE CONDENSER—E. B. LEWIS, Waterbury, Conn. Filed June 17, 1925, issued June 7, 1927.

1,631,738—VERNIER CONDENSER—F. KOCH, Newark, New Jersey. Filed April 28, 1924, issued June 7, 1927. Assigned to U. S. Tool Company, Inc.

1,631,672—VACUUM DISCHARGE APPARATUS—S. DUSHMAN, Schenectady, N. Y. Filed Mar. 4, 1919, issued June 7, 1927. Assigned to General Electric Co.

1,631,917—THERAPEUTIC OSCILLATOR—C. C. COOK and G. S. WALKER, Saginaw, Michigan. Filed Jan. 28, 1925, issued June 7, 1927.

1,632,039—ANTENNA STRUCTURE—A. A. OSWALD and E. L. NELSON, East Orange, N. J. Filed Jan. 4, 1924, issued June 14, 1927. Assigned to Western Electric Co.

1,632,054—OSCILLATION GENERATOR—J. T. L. BROWN, New Brighton, N. Y. Filed Dec. 3, 1923, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,069—WAVE TRANSMISSION SYSTEM—R. V. L. HARTLEY, South Orange, N. J. Filed Aug. 14, 1924, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,074—CONTROL APPARATUS—W. G. HOUSKEEPER, New York City. Filed Mar. 30, 1922, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,080—ELECTRIC DISCHARGE DEVICE—J. B. JOHNSON, Elmhurst, N. Y. Filed Dec. 27, 1921, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,130—CONDENSER—A. HADDOCK, East Orange, N. J. Filed Dec. 7, 1925, issued June 14, 1927. Assigned to Bell Telephone Laboratories, Inc.

1,632,135—ELECTRODE STRUCTURE FOR ELECTRON DISCHARGE DEVICES—W. F. HENDRY, New York, N. Y. Filed April 1, 1920, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,150—RADIO RECORDER—H. P. SPARKES, Edgewood Park, Pa. Filed Sept. 23, 1921, issued June 14, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,632,252—TWO-WAY WAVE TRANSMISSION SYSTEM—J. F. FARRINGTON, Flushing, N. Y. Filed Dec. 29, 1923, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,369—RADIO SIGNALING SYSTEM—A. CROSSLEY, Washington, D. C. Filed Oct. 12, 1925, issued June 14, 1927. Assigned to Wired Radio, Inc.

1,632,389—OSCILLATION GENERATING AND SUPPLY SYSTEM—A. M. CURTIS, East Orange, N. J. Filed Dec. 8, 1923, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,487—VARIABLE CONDENSER—R. H. MANSON, Rochester, N. Y. Filed Feb. 16, 1923, issued June 14, 1927. Assigned to the Stromberg-Carlson Telephone Mfg. Co.

1,632,488—VARIABLE CONDENSER—R. H. MANSON, Rochester, N. Y. Filed Nov. 24, 1924, issued June 14, 1927. Assigned to the Stromberg-Carlson Telephone Mfg. Co.

1,632,615—VACUUM TUBE—L. E. MITCHELL and A. J. WHITE, East Cleveland, Ohio. Filed June 5, 1924, issued June 14, 1927. Assigned to General Electric Co.

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1,632,649—VARIABLE CONDENSER—F. O. HARTMAN, Mansfield, Ohio. Filed Nov. 23, 1925, issued June 14, 1927. Assigned to Hartman Electrical Mfg. Co.

1,632,650—VARIABLE CONDENSER—F. O. HARTMAN, Mansfield, Ohio. Filed Dec. 8, 1925, issued June 14, 1927. Assigned to Hartman Electrical Mfg. Co.

Re. 16,651—ADJUSTMENT OF WAVELENGTHS—G. H. CLARK, Washington, D. C. Filed (original) Jan. 18, 1915; reissue filed April 20, 1920, issued June 14, 1927. Assigned to Radio Corp. of America.

1,632,870—THERMIONIC VALVE—A. C. BARTLETT and R. LE ROSSIGNOL, of Waterford, and Harrow, England, respectively. Filed Nov. 17, 1926, issued June 21, 1927. Assigned to General Electric Co. Ltd.

1,632,878—RADIO CONDENSER COUPLING—H. A. BREMER, Chicago, Ill. Filed May 24, 1926, issued June 21, 1927.

1,632,982—WAVE METER—J. O. MAUBORGNE and GUY HILL, of Washington, D. C. Filed Aug. 6, 1920, issued June 21, 1927.

1,633,019—ELECTRICAL CONDENSER—W. H. HUTH, Chicago, Ill. Filed Nov. 19, 1925, issued June 21, 1927.

1,633,059—CRYSTAL DETECTOR—H. C. ADAM, St. Louis, Missouri. Filed Mar. 12, 1923, issued June 21, 1927.

1,633,100—PLURAL CHANNEL SIGNALING—R. S. HEISING, East Orange, N. J. Filed Mar. 3, 1916, issued June 21, 1927. Assigned to Western Electric Co., Inc.

1,633,285—HIGH POTENTIAL ELECTRICAL CONDENSER—J. A. PROCTOR, Lexington, Mass. Filed Feb. 1, 1921, issued June 21, 1927. Assigned to Wireless Specialty Apparatus Co.

1,633,481—FREQUENCY REDUCING ARRANGEMENT—J. G. FALLOU, Paris, France. Filed Nov. 8, 1926, issued June 21, 1927. Assigned to L'Union Generale D'Electricite.

1,633,516—MULTIPLE VARIABLE CONDENSER—W. M. BROWER, Palo Alto, Calif. Filed Jan. 17, 1927, issued June 21, 1927. Assigned to Federal Telegraph Co.

1,633,566—ELECTRICAL CONDENSER—W. A. BOCKIUS, of Park Ridge, Ill. Filed Feb. 13, 1925, issued June 28, 1927.

1,633,775—ANTENNA ARRANGEMENT FOR SHORT-WAVE APPARATUS—A. ESAU, of Jena, Germany. Filed Oct. 8, 1926, issued June 28, 1927.

1,633,870—CONDENSER—E. F. POTTER, of Glencoe, Ill. Filed Jan. 26, 1923, issued June 28, 1927. Assigned to Kellogg Switchboard and Supply Co.

1,633,915—CONDENSER—G. A. YANOCHOWSKI, of Chicago, Ill. Filed Jan. 25, 1923, issued June 28, 1927. Assigned to Kellogg Switchboard and Supply Co.

1,633,932—RADIO RECEIVING SYSTEM—F. B. FALKON, of Wilkinsburg, Pa. Filed April 26, 1923, issued June 28, 1927. Assigned to Westinghouse Electric & Manufacturing Company.

1,634,390—RADIO TRANSMITTING SYSTEM—V. K. ZWORYKIN, of Wilkinsburg, Pa. Filed March 17, 1924 issued July 5, 1927. Assigned to Westinghouse Electric and Mfg. Co.

1,634,627—SPARK GAP—H. H. OSBORN, of Chicago, Ill. Filed June 10, 1921, issued July 5, 1927. Assigned to H. G. Fischer & Co.

1,634,896—RADIO APPARATUS—L. B. COGSWELL, of West Springfield, Mass. Filed March 25, 1926, issued July 5, 1927.

1,634,930—VARIABLE CONDENSER—C. P. BROCKWAY, of Toledo, Ohio. Filed Nov. 22, 1923, issued July 5, 1927. Assigned to Arcturus Radio Co.

1,634,962—RADIO RECEIVING SYSTEM—R. SACHTLEBER, of Brooklyn, New York. Filed Sept. 22, 1924, issued July 5, 1927.

1,635,117—SIGNAL RECEIVING SYSTEM—F. W. DUNMORE, of Washington, D. C. Filed Feb. 27, 1922, issued July 5, 1927.

1,635,152—RADIO BROADCAST SELECTING AND DISTRIBUTING SYSTEM—EDWARD E. CLEMENT, of Washington, D. C. Filed Feb. 29, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,153—SUBDIVIDED SERVICE SYSTEM OF RADIO BROADCAST DISTRIBUTION—E. E. CLEMENT of Washington, D. C. Filed Oct. 28, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

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1,635,156—RADIO BROADCAST DISTRIBUTING SYSTEM—E. E. CLEMENT, of Washington, D. C. Filed Oct. 21, 1925, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,157—RADIO BROADCAST DISTRIBUTING SYSTEM—E. E. CLEMENT, of Washington, D. C. Filed Oct. 28, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,158—RADIO BROADCAST DISTRIBUTING SYSTEM—E. E. CLEMENT, of Washington, D. C. Filed Oct. 28, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,459—RADIO CONDENSER—P. A. CHAMBERLAIN, of Chicago, Ill. Filed Sept. 18, 1925, issued July 12, 1927. Assigned to Mohawk Electric Corp.

1,635,556—CONDENSER—R. E. MARBURY, of Wilkinsburg, Pa. Filed Jan. 16, 1924, issued July 12, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,635,990—ELECTROSTATIC CONDENSER—JOHN O. GARGAN, Brooklyn, N. Y. Filed Aug. 26, 1924, issued July 19, 1927. Assigned to Western Electric Co., Inc.

1,635,992—ELECTRON DISCHARGE DEVICE—A. HADDOCK, East Orange, N. J. Filed Oct. 4, 1924, issued July 19, 1927. Assigned to Western Electric Co., Inc.

1,635,999—ELECTRON DISCHARGE DEVICE—W. G. HOUSKEEPER, New York, N. Y. Filed Sept. 22, 1922, issued July 19, 1927. Assigned to Western Electric Co.

1,636,015—ELECTRON DISCHARGE DEVICE—V. L. RONCI, Brooklyn, N. Y. Filed April 27, 1923, issued July 19, 1927. Assigned to Western Electric Co., Inc.

1,636,094—ARC GENERATOR—R. HERZOG, Berlin, Germany. Filed Dec. 29, 1925, issued July 19, 1927. Assigned to C. Lorenz Aktiengesellschaft.

1,636,239—VACUUM TUBE—H. E. METCALF, San Leandro, Calif. Filed Oct. 13, 1924, issued July 19, 1927. Assigned to The Magnavox Company.

1,636,261—VACUUM OR THERMIONIC TUBE—A. G. THOMAS, Lynchburg, Virginia. Filed Aug. 2, 1923, issued July 19, 1927.

1,636,269—CRYSTAL DETECTOR—E. A. WRIGHT, Chicago, Ill. Filed Aug. 22, 1925, issued July 19, 1927. Assigned to F. H. Noble Co.

1,636,328—CONDENSER—L. B. SAUER, Western Springs, Ill. Filed May 23, 1924, issued July 19, 1927. Assigned to Kellogg Switchboard and Supply Co.

1,636,503—VARIABLE CONDENSER—H. G. DORSEY, Gloucester, Mass. Filed Nov. 11, 1922, issued July 19, 1927. Assigned to Submarine Signal Company.

1,636,570—RADIO RECEIVING APPARATUS—F. A. KOLSTER, Palo Alto, Calif. Filed Jan. 24, 1926, issued July 19, 1927. Assigned to Federal Telegraph Co.

1,636,921—PIEZO AUDION—A. McL. NICOLSON, New York City. Filed Nov. 11, 1926, issued July 26, 1927. Assigned to Wired Radio, Inc.

1,637,045—RADIO RECEIVING SYSTEM—EDWARD W. KELLOGG, Schenectady, N. Y. Filed April 15, 1922, issued July 26, 1927. Assigned to General Electric Co.

1,637,119—ELECTROMAGNETIC SOUND REPRODUCER—F. A. KOLSTER and S. A. SOLLIE. Filed Jan. 18, 1927, issued July 26, 1927. Assigned to Federal Telegraph Co.

1,637,310—TRANSMISSION SYSTEM FOR RADIANT ENERGY—J. H. HAMMOND, JR., Gloucester, Mass. Filed originally Sept. 28, 1917, renewed Aug. 1, 1924 and issued July 26, 1927.

Des. 72,871—CONE SPEAKER—M. C. RYPINSKI, L. W. STAUNTON and S. BOURNE, of Yonkers, N. Y.; Jackson Heights, N. Y.; and New York City, respectively. Filed Mar. 10, 1926, issued June 14, 1927. Assigned to Brandes Laboratories, Inc.

1,637,119—ELECTROMAGNETIC SOUND REPRODUCER—F. A. KOLSTER and S. A. SOLLIE, of Palo Alto, Cal. Filed Jan. 18, 1927, issued July 26, 1927. Assigned to Federal Telegraph Company.

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	Rochester, 72 Bedford St.,	Miller, R. G.
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	N. Auckland, Parenga, Cape Maria Radio,	Tyler, G. E.

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 15 SEPTEMBER, 1927 Number 9

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OBITUARY

The Institute learns with regret of the death on August 16th, 1927 of

Charles V. Logwood

Mr. Logwood, who was born in California in 1882, has long been identified with the development of continuous wave radio reception and the vacuum tube. He was an active worker in the art, and the grantee of a number of radio patents. The sudden curtailment of his efforts is a distinct loss to the radio art and industry.

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THE CORRELATION OF RADIO RECEPTION WITH SOLAR ACTIVITY AND TERRESTRIAL MAGNETISM. II.*

By

GREENLEAF W. PICKARD

(Consulting Engineer, the Wireless Specialty Apparatus Company, Boston, Massachusetts.)

(Communication from the International Union of Scientific Radio Telegraphy.)

This paper is a continuation of an earlier one[1] on the same subject, in which day reception at 15.25 kilocycles, night reception at 1330 kilocycles and night reception at 8-9 megacycles were compared with sunspot numbers and magnetic measures. Although in the former paper definite relations were shown, there were also discrepancies, particularly in the relation of sunspots to reception and magnetism. When the three elements were compared over any considerable period, the maxima and minima swung alternately in and out of step in an irregular seeming manner, although preserving in a general way a 27-day period.

It must be confessed at the outset that the relation between individual sunspots, reception and magnetism is still far from definite. In general it appears that the terrestrial elements are most disturbed when the sunspots are large and numerous, and particularly when they face most nearly earthward. But often large spots cross the meridian without either accompanying magnetic disturbances or changes in reception, and it is well-known that severe magnetic storms sometimes occur when the only sunspots visible are near the sun's limbs.

Undoubtedly both magnetism and reception are subject to disturbances which are not of solar origin, or which at

*Presented before the American Geophysical Union, Section of Terrestrial Magnetism and Electricity, Washington, D. C., April 28, 1927.

Received by the Institute April 20, 1927.

[1]"The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism, Proceedings of the Institute of Radio Engineers, Vol. 15, No. 2, February, 1927.

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least do not follow either solar rotation periods or the longer swing of the sunspot cycle. Thus, diurnal periods are found in both reception and magnetism, and seasonal changes are also well known. A solar eclipse distinctly affects radio reception[3] and less definitely disturbs terrestrial magnetism. There also appear to be relations between certain meteorological elements and reception which are not paralleled by magnetic disturbances, probably

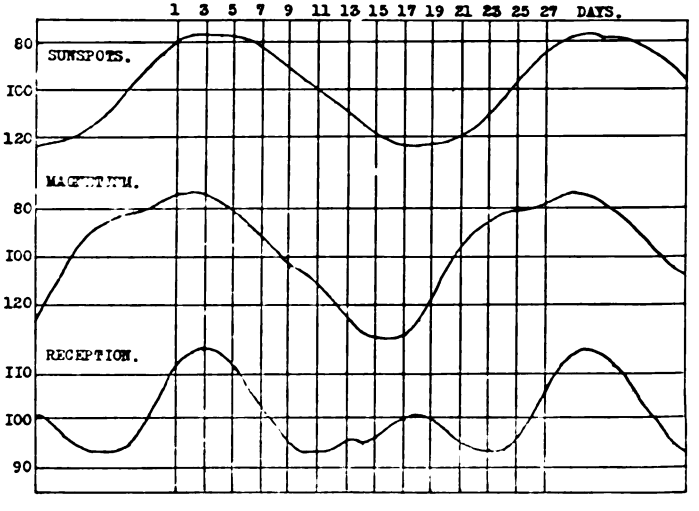

Fig. 1—Mean of 12 Solar Rotations, October 25, 1908 to September 17, 1909. Sunspots, Magnetic Character of Day and Maximum Night Ranges of "DF" Working.

because reception and weather are rather local matters, while terrestrial magnetism is quite literally a world-wide affair.

While reception disturbances of non-solar periods are important and fertile fields for investigation, the present paper must largely confine itself to magnetic and reception changes associated with the solar rotation period, and so these other effects must be eliminated rather than analyzed. This is readily done by taking a sufficiently long series of observations, dividing this series into 27.3 day periods, and finding the daily means. This operation averages out anything which does not recur in each rotation, and gives a

[3]The Effect of the Solar Eclipse of January 24, 1925, on Radio Reception. Proceedings of the Institute of Radio Engineers, Vol. 13, No. 5, October, 1925.

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clearer picture of the interrelation than that presented in the former paper.

It is not necessarily true that the solar centers responsible for terrestrial disturbances are the sunspots themselves, although the evidence is now strong that these centers are at least closely associated with sunspots or sunspot groups. If it be assumed that these centers, like

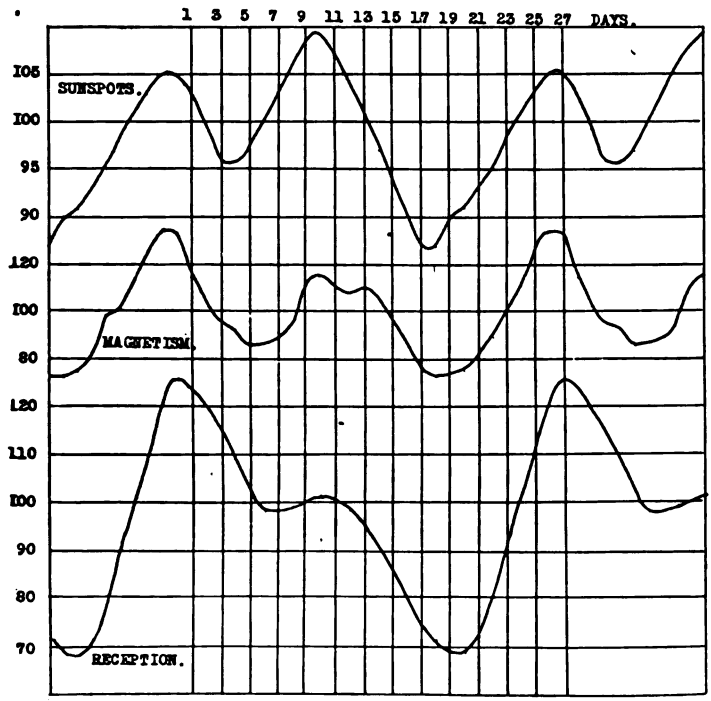

Fig. 2—Mean of 5 Solar Rotations, January 14 to May 30, 1917. Sunspots, Magnetic Character of Day and Washington Day Reception of POZ.

the accompanying sunspots, are confined within definite though rather wide zones for long periods, but shift about within these areas from month to month, some of the discrepancies found can be quite simply explained. And even if, as at present appears, the wanderings of the active centers are entirely at random, the mean of a number of solar rotations should show definite disturbance maxima corresponding with the meridian crossing of the centers of the active zones, instead of the somewhat irregularly spaced

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disturbances accompanying the transit of the individual centers.

Before periodic means of sunspot numbers, magnetic measures, or reception observations are made, it is usually necessary to reduce the numerical values to such a basis that periods of high absolute values will not dominate the means of a long series. This is easily done by converting the daily values into ratios with a moving mean of the

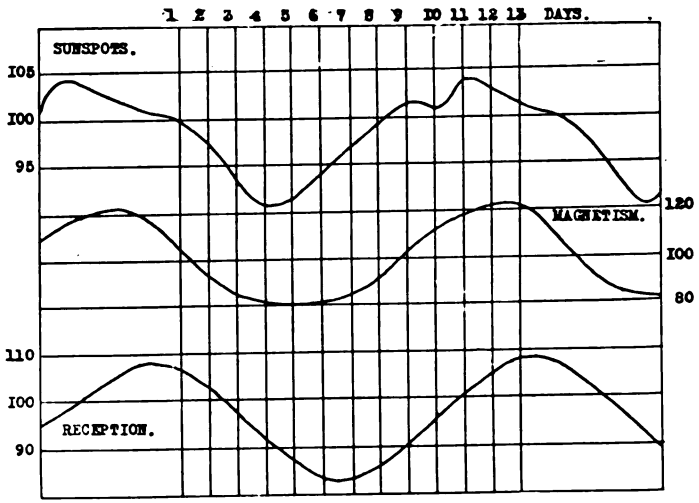

Fig. 3—Mean Within 5 Solar Rotations of Ten 13.64 Day Periods, January 14 to May 30, 1917. Sunspots, Magnetic Character of Day and Washington Day Reception of POZ.

period under investigation, thus obtaining a measure of interperiod activity. By taking a sufficient number of periods, even rough measurements of reception can be made to show definite relations to sunspots and terrestrial magnetism.

In Fig. 1 the reception values are the maximum nightly ranges of two-way working between station DF at Manhattan Beach, New York, and various ships at sea[3], which are compared with the Wolfer Final Sunspot Numbers and van Dijk's Magnetic Character of Day Numbers. The interval taken is from October 25, 1908 to September 17, 1909, corresponding to 12 solar rotations of 27.3 days. It will be seen that the sunspot curve is nearly sinusoidal, indicat-

[3]Short Period Variations in Radio Reception, Proceedings of the Institute of Radio Engineers, Vol. 12, No. 2, April, 1924, pages 119-158.

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ing that in the sunspot numbers the only frequency of appreciable amplitude is that of the solar rotation, but both magnetic and reception curves show the presence of shorter periods. Owing to the roughness of the reception data, no attempt has been made to investigate the harmonics of the solar period, although this has been done for some of the later and more accurate reception measurements. The

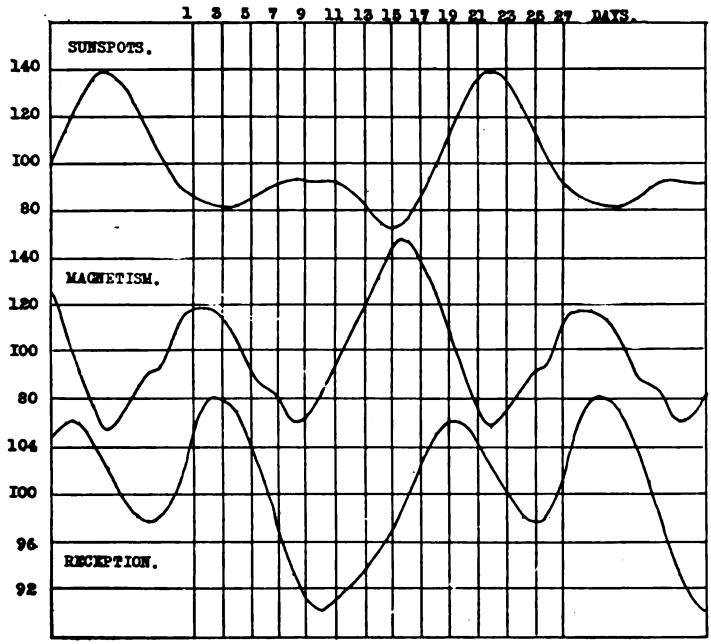

Fig. 4—Mean of 9 Solar Rotations, March 14 to November 15, 1922. Sunspots, Magnetic Character of Day and Washington Day Reception from POZ and LY.

relation of reception to the other elements is inverse, which appears characteristic of night reception save at ultra high frequencies. It is of interest to note that this transmission was all at frequencies within the present broadcasting band, so that the results may be compared with recent measurements, allowance being made for the fact that in 1908-1909 spark transmitters of relatively high decrement were used.

During 1917, a sunspot maximum year, the U. S. Naval Radiotelegraphic Laboratory made audibility meter measurements at Washington of day reception from station POZ at Nauen, Germany. Owing to war conditions, Dr.

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Austin considers these the least accurate of his long series of measurements of this station; nevertheless Fig. 2 clearly shows relation between this reception, sunspots and

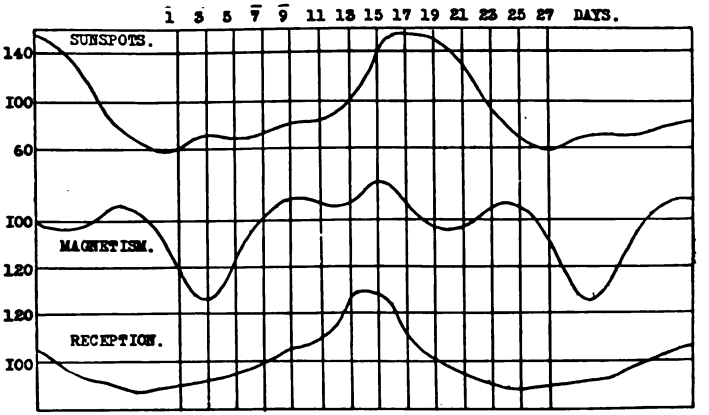

Fig. 5—Mean of 7 Solar Rotations, May 14 to November 21, 1923. Sunspots, Magnetic Character of Day and Washington Day Reception of POZ.

magnetism. All three curves show shorter periods of considerable amplitude, and despite the roughness of the reception measurements, an analysis has been attempted for the second harmonic of the solar rotation period, or 13.6 days. While in Fig. 1 the relation of night reception to sunspots

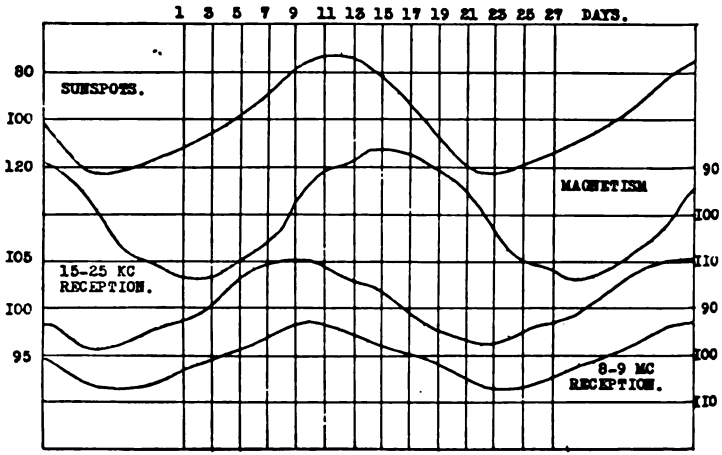

Fig. 6—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham, 15—25 Kilocycle Day Reception at Washington and 8—9 Megacycle Night Reception at Washington.

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and magnetism is inverse, in Fig. 2 reception is found to be directly correlated with the other elements; this seems to be the normal relation of day reception at low frequency.

In Fig. 3 means of 1917 day reception, sunspots and magnetism are taken for periods of 13.6 days, or the second harmonic of the solar rotation period. Considering the roughness of the reception data, the curves are in excellent agreement, and their nearly sinusoidal form indicates that there are no important shorter periods present. The low

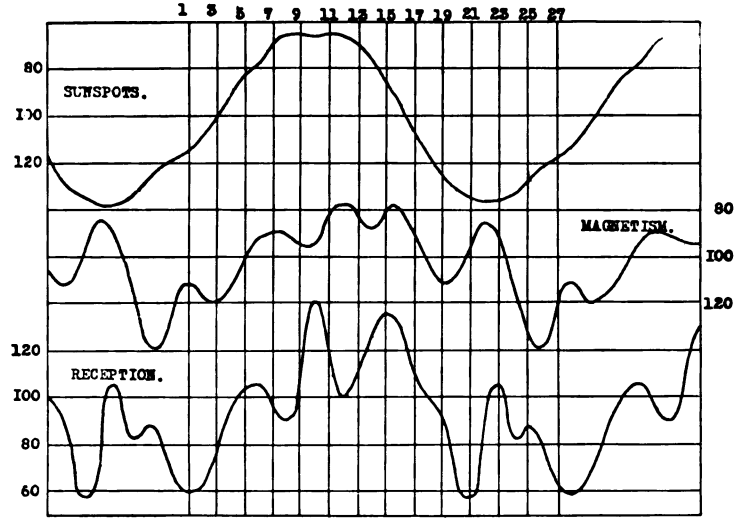

Fig. 7—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Center from WBBM at Chicago. No Smoothing.

amplitude of the sunspot curve is due to the nature of the Wolfer numbers, which are taken over the entire visible solar disk, or over nearly 180 deg. in longitude. This process leaves very little of the shorter periods, although it will be later shown that these are prominent in sunspot numbers taken in restricted zones instead of over nearly a hemisphere.

Figs. 4 and 5 are of Washington day reception for the years 1922 and 1923, the latter year being a sunspot minimum. Again the relation of day reception to solar activity and magnetism is found to be direct, and the presence of shorter and probably harmonic periods is strongly indicated. The irregularity of the 1923 sunspot curve is in part

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due to the fact that 47 per cent of the days taken had a sunspot number of zero.

In Fig. 6 a comparison is made between sunspots, diurnal range of *H* at Cheltenham, 15-25 kilocycle day reception for a group of 9 stations, and 8-9 megacycle night reception for a group of 7 stations. A close agreement between the two reception curves is evident, but the sunspot and

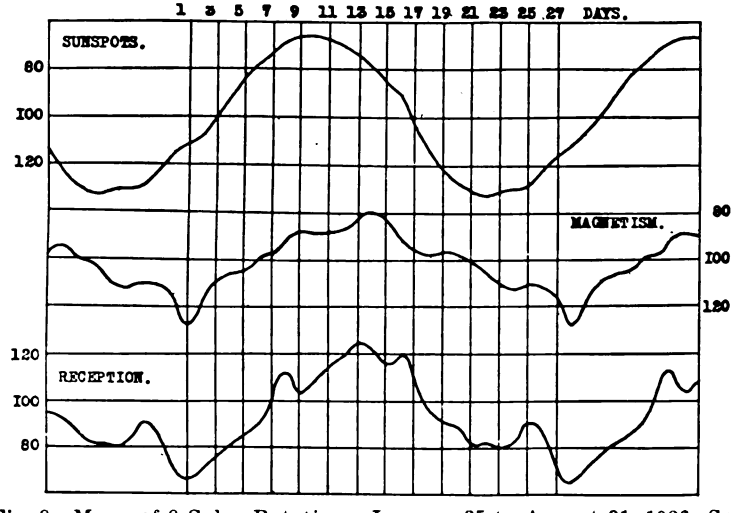

Fig. 8—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Centre from WBBM at Chicago. Smoothed by Moving 7-Day Mean.

magnetic curves lag 4 and 6 days, respectively, behind reception. The relation between sunspots, magnetism and 8-9 megacycle night reception is direct, but is inverted for 15-25 kilocycle day reception. This does not agree with the findings for other years, and apparently 1926 is an exception to a general rule. But the data have been taken from a single station in the preceding years, instead of the group used for 1926, so the apparent inversion may be subject to revision after further study of the data.

As set forth in the previous paper, it appears that reception in the frequency band of 500 to 1500 kilocycles shows higher correlation with solar activity and terrestrial magnetism than does any other investigated portion of the radio spectrum, and for that reason the author's measurements have been largely confined to night broadcast re-

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ception. In Fig. 7 are given the means within a 27.3 day period of the Wolfer Provisional Sunspot Numbers, the diurnal range of the earth's horizontal magnetic field as taken by Cheltenham Observatory and mean night field at Newton Centre, Mass., from station WBBM at Chicago, operating at 1330 kilocycles. The sunspot curve is nearly sinusoidal, indicating the absence in the Wolfer numbers of any large amplitude periods shorter than the fundamen-

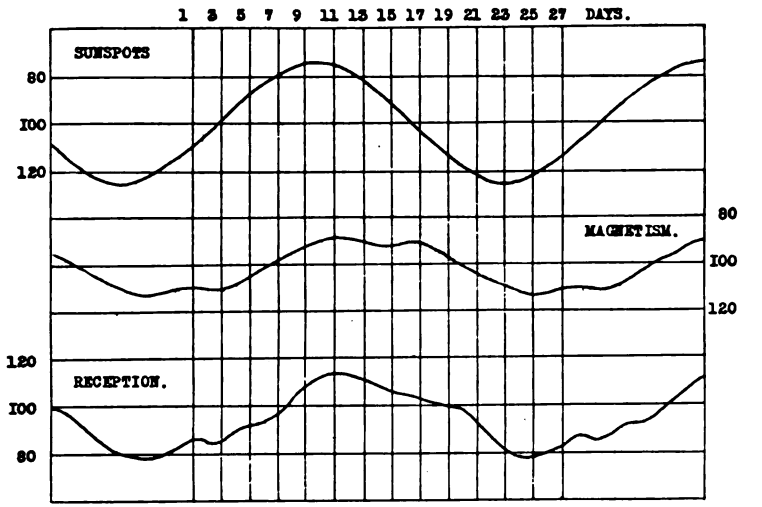

Fig. 9—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Centre from WBBM at Chicago. Smoothed by 13-day Moving Mean.

tal, but the magnetic and reception curves evidently contain appreciable shorter period components. The fundamental 27.3 day period is best shown by eliminating the shorter periods, which is partially done in Fig. 8 by smoothing with a moving 7-day mean, and more completely in Fig. 9 by a moving 13-day mean, which leaves little but the fundamental.

In Fig. 10 periodic daily means are taken within the second harmonic of the solar rotation period, or 13.6 days. This greatly reduces the amplitude of the sunspot curve, but the reception and magnetic curves still show the presence of shorter periods. In Fig. 11 the third harmonic of 9.1 days is taken, which still further reduces the amplitude of the sunspot curve, leaves the magnetic curve nearly sinusoidal, but

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the reception curve still shows a shorter period component. Finally in Fig. 12 a 6.8 or fourth harmonic period is used, which reduces the sunspot curve to very small amplitude, while the magnetic and reception curves now become quite similar, with little indication of any shorter period component.

As indicated above, the Wolfer sunspot numbers cannot be effectively used in the investigation of the shorter

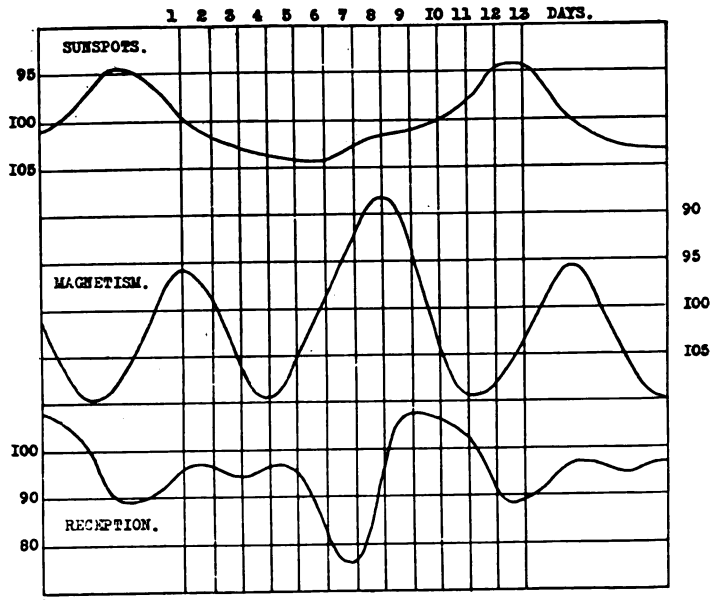

Fig. 10—Mean Within 8 Solar Rotations of Sixteen 13.64 Day Periods, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Centre of WBBM.

solar periods. Through the kindness of the U. S. Naval Observatory a list has been given of the times of meridian passage of the larger sunspots and sunspot groups of 1926, and a comparison of this with reception measures shows very clearly that night reception depressions are in general coincident with the meridian transit of the spots, while the appearance of large spots and groups near the sun's limbs had little or no correlation with reception. Through the courtesy of Dr. H. H. Clayton, who has taken daily sunspot observations for the past several years, sunspot numbers for a zone ten degrees on each side of the central solar meridian have been furnished for each day of 1926. Using these

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sunspot numbers and night reception from WBBM, the relations within the fundamental 27.3 day period, and its second, third, and fourth harmonics are given in Figs. 13 and 14. The correlation between the sunspot numbers for the central twenty degree zone and reception is now found to be very high, and the amplitude of the sunspot curve for even the fourth harmonic of 6.8 days is quite large. Not only are

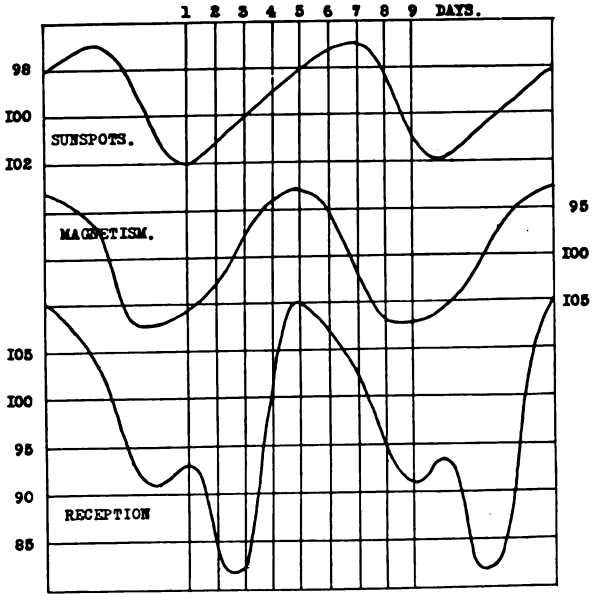

Fig. 11—Mean Within 8 Solar Rotations of Twenty-Four 9.1 Day Periods, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception of WBBM.

the curves closely alike in form, but the maxima and minima agree to the nearest day, which confirms the conclusion reached from the study of the U. S. Naval Observatory data, that reception disturbances generally coincide with the earthward presentation of the sunspots.

Measurements of night static were also taken at 1330 kilocycles, and periodic means made over 14 solar rotations, or from January 25, 1926 to February 11, 1927. The result, as compared with sunspots, is shown in Fig. 15. Unlike the night signal at the same frequency, the relation is direct. An investigation of day static in the 15-25 kilocycle band shows an inverse relation to sunspots, although the agree-

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ment is not so good as that shown in Fig. 15. At present there is no satisfactory explanation for this relation, either

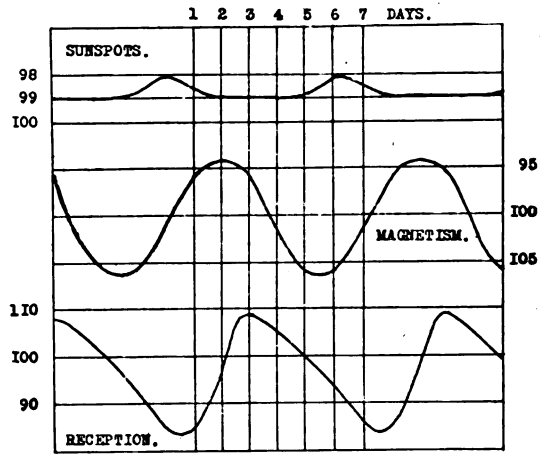

Fig. 12—Mean Within 8 Solar Rotations of Thirty-Two 6.8 Day Periods, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception from WBBM.

on a basis of static wave transmission, or the assumption of electric or other storm areas as static sources. Both in

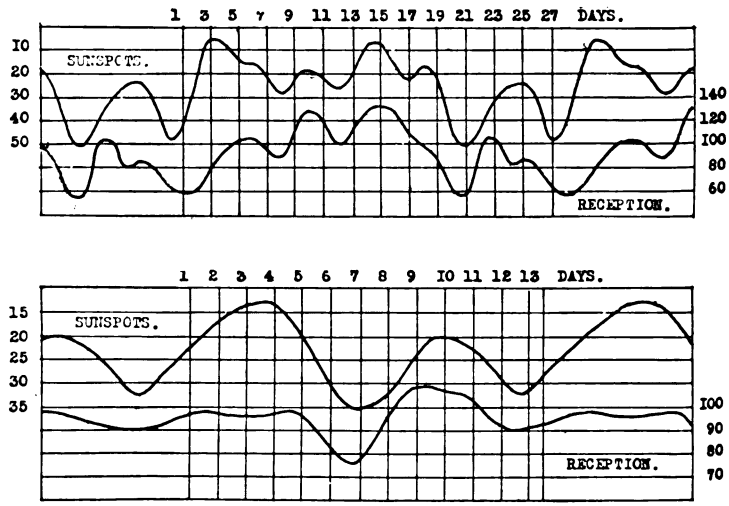

Fig. 13—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots Within Ten Degrees of Solar Meridian and Night Reception of WBBM. Fundamental and Second Harmonic of Solar Reception.

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day reception at low frequencies, and in night reception in the broadcast band, the static is inversely related to signal, so it is difficult to frame an hypothesis based on transmission, while if the relation depends upon variability of static sources it must be assumed that the activity of static centers at night is inversely related to that of the day, and at the same time both are linked with solar activity.

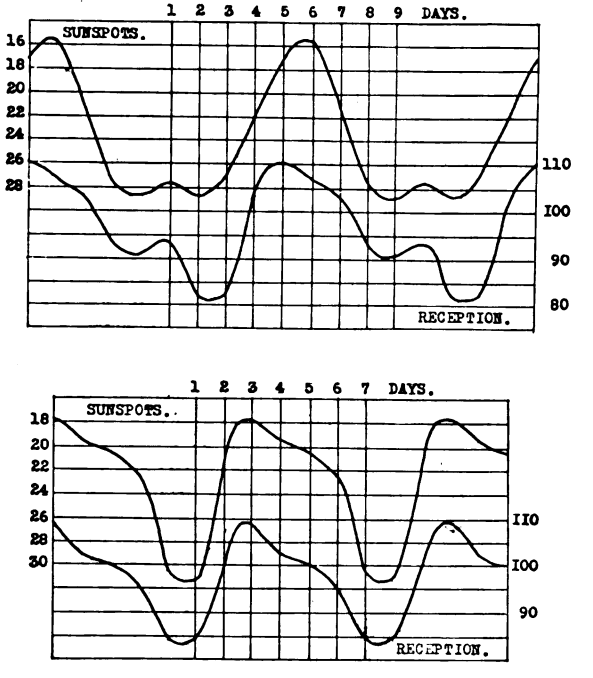

Fig. 14—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots Within Ten Degrees of Solar Meridian and Night Reception of WBBM. Third and Fourth Harmonics of Solar Rotation.

In conclusion, the existence of a pronounced annual double periodicity reception component, paralleling that already recognized in terrestrial magnetism, is shown in Fig. 16. Here periodic monthly means are made over an interval of 8 years, from 1916 to 1924, and well-defined maxima show at or near the vernal and autumnal equinoxes. The relation between magnetism and reception is direct, for the reception is by day; if such a graph could be made for night reception it would probably show minima at the equinoxes. While it will be several years before really ade-

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quate data are accumulated for night reception in the broadcast band, it may be of interest to note that a summation of night spark reception in 1908-1909, scattered observations

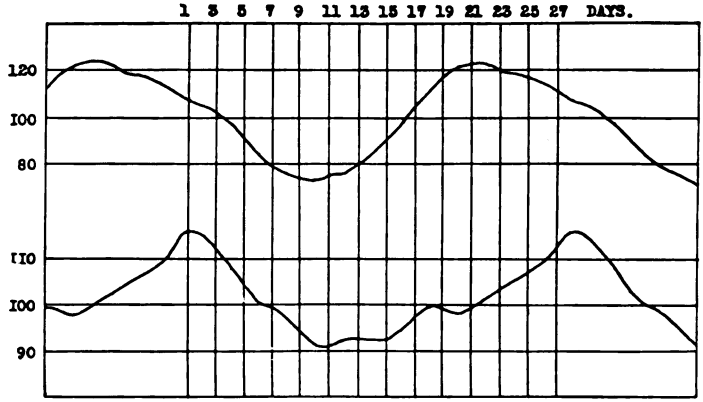

Fig. 15—Mean of 14 Solar Rotations, January 25, 1926 to February 11, 1927. Sunspots and Night Static at 1330 kilocycles.

in 1923-1924 and the WBBM measurements of 1926-1927 shows distinct minima in the spring and fall.

For the convenience of other workers, Dr. Clayton's 1926 sunspot numbers for 20 deg. and 40 deg. central zones are appended. And supplementing the table in the former

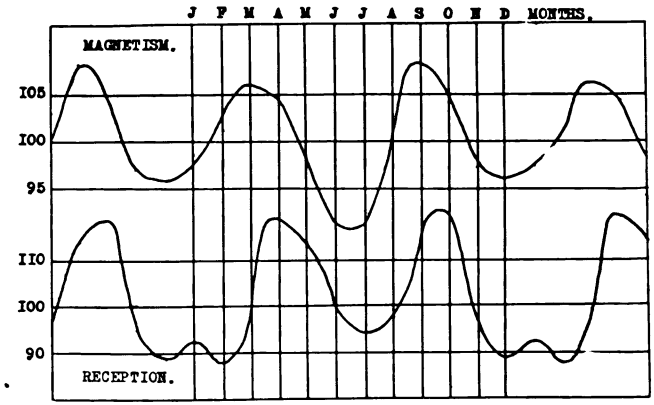

Fig. 16—Mean of 8 years, 1916 to 1924. Monthly Averages of Magnetic Character of Day and Washington Day Reception from POZ.

paper, nightly values of WBBM's field are given from December 15, 1926 to March 31, 1927.

Dr. Clayton has recently found an interesting relation

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NUMBER OF SPOTS (UMBRAE) WITHIN 20 DEG. OF CENTRAL MERIDIAN OF THE SUN. ZONE 40 DEG. WIDE.

1926

Day	Jan.	Feb.	Mar.	Apr.	May	June	July	Aug.	Sept.	Oct.	Nov.	Dec.
1	(4)	18	(15)	(0)	0	(8)	19	3	2	1	2	(14)
2	(5)	(10)	40	(0)	7	12	(5)	2	3	(1)	(3)	(12)
3	1	(3)	20	(1)	10	16	7	8	(5)	6	3	10
4	(12)	(0)	27	1	16	9	2	9	3	4	2	5
5	(14)	0	14	1	22	(3)	5	9	(5)	0	0	(7)
6	(14)	(0)	8	1	24	(2)	(7)	4	(10)	(0)	2	4
7	3	(0)	(12)	(0)	6	(4)	(4)	15	10	4	11	5
8	1	0	18	(0)	1	12	(0)	19	2	14	10	0
9	(0)	0	13	0	2	36	(0)	16	0	21	(5)	5
10	7	(0)	13	0	12	(17)	(0)	0	0	(25)	6	(17)
11	16	0	2	2	10	9	0	3	0	17	9	(12)
12	(22)	7	(12)	1	20	3	0	3	3	16	0	12
13	9	(17)	14	2	19	0	0	(1)	5	12	1	1
14	(9)	(22)	7	7	(19)	0	(0)	1	2	11	(3)	16
15	4	14	8	5	0	(7)	(0)	7	3	13	7	(29)
16	5	8	7	8	(0)	14	(0)	(7)	4	10	7	27
17	(4)	9	(8)	8	0	9	0	7	(7)	(11)	8	8
18	(1)	(10)	11	7	0	1	0	2	21	(1)	3	1
19	2	(4)	8	(5)	5	4	0	4	22	(0)	4	1
20	(9)	9	(4)	0	20	10	2	5	24	(0)	(0)	(0)
21	(22)	12	1	0	16	11	0	1	8	(2)	0	0
22	32	(8)	1	0	11	(11)	2	0	1	2	14	12
23	30	0	(1)	0	(0)	(7)	11	0	(3)	3	14	16
24	39	0	1	0	0	1	(15)	3	(9)	(4)	15	23
25	(26)	0	4	0	0	1	20	(8)	4	(7)	2	11
26	(6)	0	(1)	0	(0)	1	1	(8)	(0)	7	(3)	(0)
27	0	(0)	0	0	0	5	5	6	(3)	3	6	0
28	(0)	(0)	3	0	3	8	18	4	(6)	(3)	12	(0)
29	3		6	0	0	25	(29)	1	4	4	(10)	0
30	3		10	(0)	(2)	25	(29)	3	2	9	(9)	2
31	(12)		(7)		(4)		(16)	2	1	(7)		9

Figures in brackets taken from the days preceding or following, allowing for the rotation of the sun.

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NUMBER OF SPOTS (UMBRAE) WITHIN 10 DEG. OF THE CENTRAL MERIDIAN OF THE SUN. ZONE 20 DEG. WIDE.

1926

Day	Jan.	Feb.	Mar.	Apr.	May	June	July	Aug.	Sept.	Oct.	Nov.	Dec.
1	(1)	12	(10)	(0)	0	(2)	4	1	2	1	1	(5)
2	(2)	(0)	7	(0)	0	8	(1)	0	0	(1)	(1)	10
3	1	(0)	17	(0)	8	12	0	3	(0)	0	1	0
4	(0)	(0)	15	1	4	1	0	8	3	4	2	0
5	(17)	0	4	0	15	(0)	4	0	(2)	0	0	(7)
6	(5)	(0)	(0)	0	12	(0)	(2)	0	(7)	(0)	1	5
7	0	(0)	1	(0)	2	(1)	(3)	2	8	0	10	0
8	0	0	6	(0)	1	2	(0)	19	0	11	3	(1)
9	0	0	13	0	0	26	(0)	4	0	4	(0)	7
10	0	(0)	6	0	3	(19)	0	0	0	7	6	(5)
11	7	0	0	1	18	2	0	2	0	14	0	(11)
12	(11)	0	(1)	1	3	2	0	2	0	0	0	5
13	6	(8)	13	0	17	0	0	0	4	12	0	(0)
14	(2)	(17)	2	1	3	0	(0)	4	1	4	1	0
15	2	8	5	5	0	(0)	(0)	(7)	2	0	6	(28)
16	3	4	6	2	0	8	(0)	3	4	10	7	15
17	1	8	(3)	3	0	6	0	0	(0)	(3)	6	0
18	(0)	(1)	6	5	0	1	0	0	7	(1)	3	1
19	0	(1)	4	(0)	0	4	0	4	21	0	0	0
20	(2)	5	(1)	0	18	0	0	0	10	(0)	(0)	(0)
21	(16)	11	0	0	15	9	(2)	0	0	(0)	0	(0) (2)
22	21	(3)	1	0	0	(13)	0	0	1	2	3	2
23	12	0	(1)	0	0	(0)	6	0	(1)	0	14	13
24	32	0	0	0(3)	0	0	(11)	0	(4)	(0)	12	20
25	(3)	(0)	4	0	0	2	0	(2)	1	(4)	0	0
26	(0)	0	(0)	0	0	0	1	(8)	(0)	3	(4)	(0)
27	0	(0)	0	0	0	0	4	0	(0)	(0)	2	0
28	(0)	0	0	0	1	0	0	1	(0)	0	4	(0)
29	2		3	0	0(3)	18	(13)	0	1	1	(9)	0
30	3		7	0	0	15	(16)	0	1	5	(3)	0
31	(0)		(0)		(0)		0	0		0		5

(1) No observation; number is from preceding or following day for spots which would have been within the central meridian zone on that date.

0 (3) Means that 3 spots developed on that meridian one day later.

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Mean field in microvolts per meter at Newton Centre, Mass., from WBBM, Chicago, 9-10 P. M., December 15, 1926, to March 31, 1927.

Day	Dec.	Jan.	Feb.	Mar.
1	. . .	2.0	1.0	0.1
2	. . .	3.2	2.4	1.0e
3	. . .	2.0e	5.2	0.3
4	. . .	1.5	7.3	3.0
5	. . .	0.7	5.3	1.5
6	. . .	8.0	8.0	0.7
7	. . .	1.3	1.0e	4.0e
8	. . .	2.2	10.0	5.8
9	. . .	4.0e	0.3	1.9
10	. . .	2.0e	0.8	3.3
11	. . .	4.0e	1.2	1.3
12	. . .	8.0e	2.6	7.8
13	. . .	3.0	0.3	4.2
14	. . .	2.9	5.0m	4.0e
15	3.6	7.9	9.6	2.9
16	2.0	3.3	2.1	0.1
17	3.5	4.0e	3.2	0.1
18	1.5	9.7	1.7	0.7
19	1.4	3.7	1.7	0.1
20	1.0e	10.0	1.6	0.7
21	10.6	10.8	16.0e	0.8
22	12.5	14.7	12.8	0.8
23	0.3	12.3	10.5	3.0
24	2.0	4.0e	0.8	2.0
25	1.9	1.7	1.9	16.0
26	8.2	1.6	6.2	0.5
27	1.0e	4.1	10.7	2.9
28	1.9	4.0e	5.8m	3.2m
29	0.7	5.5		3.5
30	2.7	7.6		0.4
31	6.5	4.0e		0.5
Means	6.0	5.0	4.8	2.5

between night reception of WBBM and the number of sunspots in a central zone of 13.3 deg. (the width of a day's travel in solar rotation). When the number of spots in this zone was under 12, the relation between sunspot number and reception was inverse; when the spot number was over 12, a direct relation was found. The predominant inverse relation which has heretofore been found for night reception is in accord with this, for in two-thirds of the cases examined for 1926, the sunspot number for the zone was under 12. The relation between sunspot number in a central zone, day reception and magnetic measures is now under investigation, and will form the subject matter of another paper.

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SUMMARY

The solar, magnetic and reception data from the author's recent paper on this subject are now compared by periodic means, and clearer pictures of their interrelations are given. The solar rotation period and its second, third, and fourth harmonics (27.3, 13.6, 9.1 and 6.8 days) are found for all three elements, the general relation being that sunspots on the solar meridian coincide with disturbances of terrestrial magnetism, lowered night reception and higher day reception. Night static at 1330 kilocycles is found inversely related to night signal reception at the same frequency and therefore directly correlated with sunspots. Day static at 15-25 kilocycles is also, although less definitely, inversely related to the day signal at the same frequency, and therefore in general inversely correlated with sunspots. Periodic means of eight years of day reception show a marked double frequency annual component, with maxima near the vernal and autumnal equinoxes, closely paralleling the well-known annual variation in terrestrial magnetism. Tables of daily sunspot numbers for 20 deg. and 40 deg. central zones are given for 1926, and night reception values for WBBM are continued from the former paper to March 31, 1927.

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THE TESTING OF AUDIO-FREQUENCY TRANSFORMER-COUPLED AMPLIFIERS

BY
H. DIAMOND[1] and J. S. WEBB[2]

INTRODUCTION

Any stage of audio-frequency amplification may properly be considered satisfactory if a speech signal impressed upon its input circuit is exactly reproduced, on a larger scale, in its output circuit. For the particular case of transformer coupling, deviation from a faithful reproduction of the applied signal is generally due to one or more of the following four effects:

1. The component frequencies constituting the signal are non-uniformly amplified, causing what may be termed as amplitude distortion.

2. The component frequencies are each shifted in phase while passing through the amplifier, the magnitude of the phase shift angle varying with the frequency of the component considered. The relative phase relationship between the various frequencies making up the reproduced signal is consequently not the same as that between the corresponding components in the original speech signal. This may be considered as causing phase distortion.

3. Harmonics are introduced in the output signal due to the tube characteristics as well as the magnetization characteristics of the iron circuit of the transformer.

4. "Howling" may occur due to the self-generation of audio-frequency oscillations because of circuit instability.

Strictly speaking, the exact performance of a transformer-coupled amplifier is accurately known only when all four of the above effects have been experimentally investigated. For the purpose of the comparison of coupling transformers, however, it is usually sufficient to determine

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quantitatively the degree of amplitude and phase distortion present, making certain qualitatively that the other two effects do not occur in sufficient magnitude to destroy the usefulness of the amplifier. The performance of a given transformer in association with a given tube may therefore be represented graphically by a curve of the type shown in Fig. 1[1], which indicates the variation of amplitude with frequency and, in addition, the dependence of the phase shift

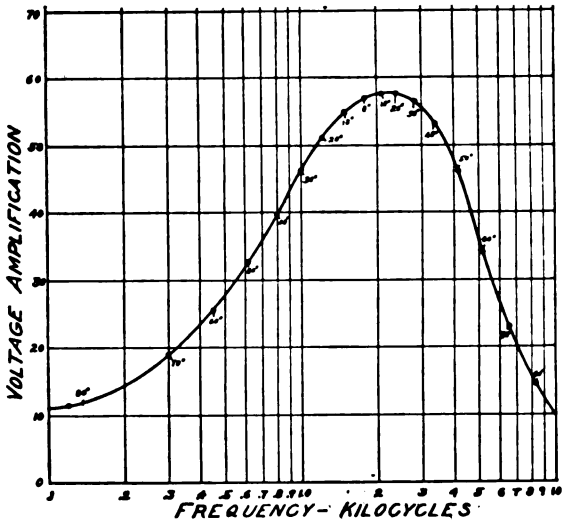

Figure 1—Typical Amplifier Performance Characteristic

angle upon the frequency. Omitting the phase shift angles gives the usual amplitude-frequency characteristic.

The curve of Fig. 1 is of little value, however, unless the conditions of operation under which the amplifier is tested are specified. The dependence of the amplifier performance upon the conditions of its use may be readily understood from a consideration of Figs. 2, 3, and 4. In Fig. 2, an alternating voltage E_{g1} of fixed frequency is applied to the grid of tube A resulting in a fictitious e.m.f. μE_{g1} in its plate circuit. Of this voltage μE_{g1}, a fraction $r_p I_p$ is consumed in the internal plate resistance of the tube and the vector remainder $E_{A\,B}$ applied to the transformer primary. This is then amplified very approximately in the ratio of turns giving the

[1]See bibliography.

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voltage E_{g2} applied to the grid of tube B. The stage amplification may be defined simply as the vector ratio E_{g2}/E_{g1}.

Fig. 3[2] is the approximately equivalent circuit of Fig. 2: —r_p represents the internal plate resistance of tube A; R_1 and X_L, the d.-c. primary resistance and the total leakage reactance, respectively; M the mutual impedance between the transformer windings exclusive of capacity; and C the

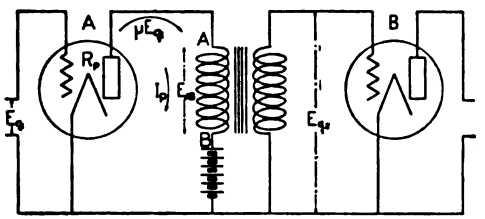

Figure 2—Circuit Diagram for a Transformer-Coupled Amplifier

combined capacitive reactance including the secondary distributed capacity, capacity between windings, etc.; all referred to the primary circuit. The impedance-frequency characteristic of the circuit of Fig. 3, to the right of points A-B, is shown in Fig. 4. This curve was obtained by actual bridge measurements on one of the transformers tested. The

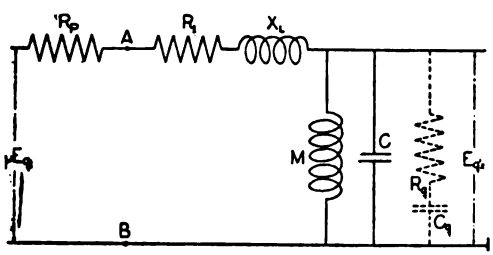

Figure 3—Equivalent Circuit for a Transformer-Coupled Amplifier

impedance Z_{AB} is seen to pass through two resonance points, the first when M and C are in parallel resonance, and the second when X_L is in series resonance with the circuit MC. The ratio E_{g2}/E_{g1} will therefore pass through a maximum value at the first resonance frequency, since Z_{AB} is then very large as compared with r_p; while it may or may not have a second maximum value at the second resonance frequency, depending on whether or not the rise in voltage across C due to

[2]See bibliography.

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series resonance is sufficient to compensate for the reduction of the impedance Z_{AB} relative to r_p.

In the above we have neglected the effect of the input impedance of tube B. In everyday use, the tube B will have in its plate circuit a complex impedance, such as another transformer, telephones, loudspeaker, etc. The reaction effect due to this load in the plate circuit may be represented by an input impedance connected between the grid and fila-

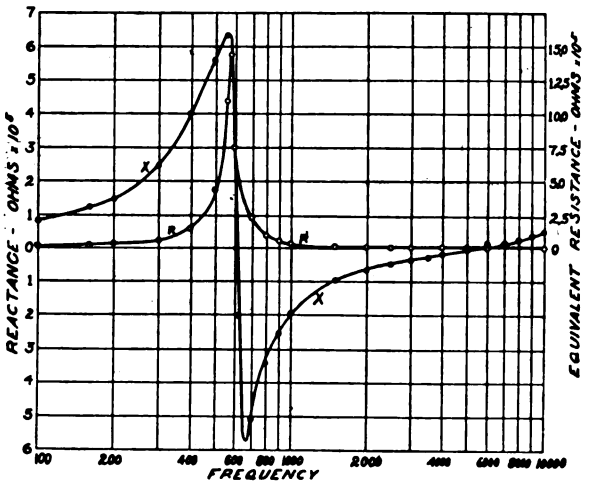

Figure 4—Transformer Impedance-Frequency Characteristic

ment of B and consisting of a resistance R_g and a capacity C_g in series.[1] The sign of this resistance is positive when the load in the plate circuit is resistive or capacitive and negative when the load is inductive. Under normal conditions of use, therefore, there will be a load across the secondary of the test transformer which varies with the frequency. The transformer performance will therefore be altered; C_g causing both resonance points to occur at lower frequencies, and R_g decreasing or increasing the overall amplification depending on its sign.

Since both the internal plate resistance of Tube A and the input impedance of tube B are seen to affect the amplifier performance, the importance of testing under normal operating conditions need hardly be stressed. In the first section of this paper it is proposed to describe several meth-

[1] See bibliography.

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ods of test whereby the performance of a given transformer and tube is obtained under actual conditions of use, no matter how varied. Results will be given showing the effect of such variation. In the second section an oscillographic method of test will be described enabling a much more rapid and visual determination of the same results. Being essentially a voltmeter method, however, the degree of phase shift cannot be measured by its means.

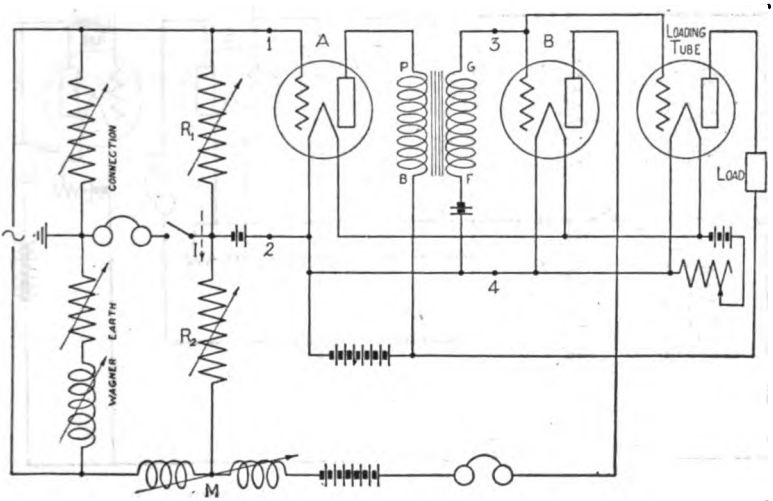

Figure 5—Circuit Arrangement for Measuring Amplification and Phase Shift.

SECTION 1.—THE MEASUREMENT OF EFFECTIVE STAGE AMPLIFICATION AND PHASE SHIFT ANGLE

Two alternative circuit arrangements for measuring the stage amplification and angle of phase shift are shown in Figs. 5 and 6. Both are essentially bridge circuits in which the amplifier output e.m.f. is balanced against an auxiliary voltage, the magnitude and phase relationship of which (relative to the amplifier input voltage) is known. In Fig. 5, this balance is effected in the plate circuit of the auxiliary tube *B*, while in Fig. 6 the balancing takes place in its grid circuit. The operation of these circuits may best be explained by means of their vector diagrams. Three such diagrams are given in Fig. 7 illustrating the operation of the circuit of Fig. 5; (a) corresponding to a frequency below the first resonance point, (b) to a frequency between the

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first and second resonance points, and (c) to a frequency above the second resonance point. Fig. 8 shows a similar set of vector diagrams for the circuit of Fig. 6.

Referring to Fig. 7 (a), $E_{g1} = IR_1$ is the voltage impressed upon the grid circuit of the amplifier tube A, $\mu_A IR_1$ the fictitious e.m.f. in the plate circuit of A and I_p the resultant plate current lagging $\mu_A IR_1$ since the equivalent impedance of the coupling transformer (referred to the

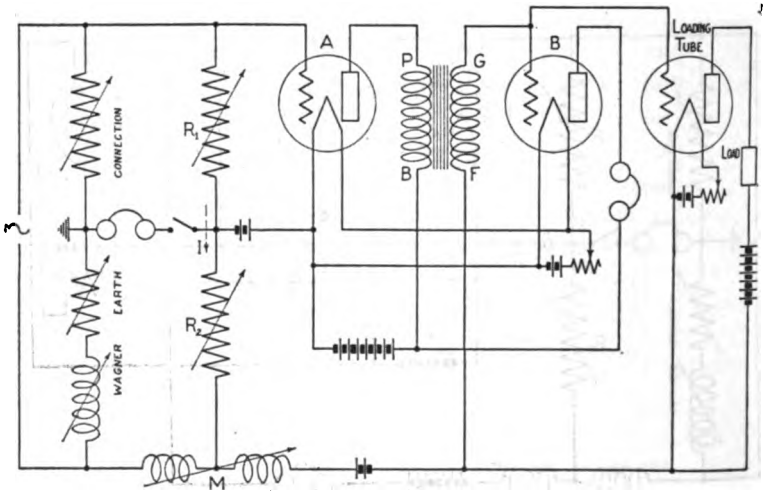

Figure 6—Alternative Method of Test

primary side) is inductive at the frequency considered. The current I_p flowing through the impedance MC results in the voltage drop E_{g2}', which when referred to the secondary side becomes $E_{g2} = nE_{g2}'$ where n is the transformer ratio of turns. This voltage E_{g2} is now impressed between the grid and filament of tube B producing the e.m.f. $\mu_B E_{g2}$ in its plate circuit. Against this e.m.f. is balanced the voltage E_2 made up of IR_2 and IwM. By adjusting R_2 and M a condition of balance may be obtained in the phones, indicating that E_2 is equal and in exact opporition to $\mu_B E_{g2}$. The amplification per stage is then given by the expression

$$K = \frac{\sqrt{(R_2)^2 + (wM)^2}}{\mu_B R_1} \qquad (1)$$

and the angle of phase shift by the expression

$$a_0 = tan^{-1} \frac{wM}{R_2} \qquad (2)$$

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The amplification factor μ_B may readily be determined by disconnecting the apparatus between leads 1-2 and 3-4, and connecting 1 to 3 and 2 to 4, respectively. The circuit is then Miller's dynamic method for measuring amplification constants.

The vector diagram of Fig. 8 (a) differs from that of 7 (a) only in that the voltage E_2 is balanced against E_{g2} rather than $\mu_B E_{g2}$. The expression for the stage amplification is therefore

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{R_1} \qquad (3)$$

the phase shift angle being as before, defined by equation (2).

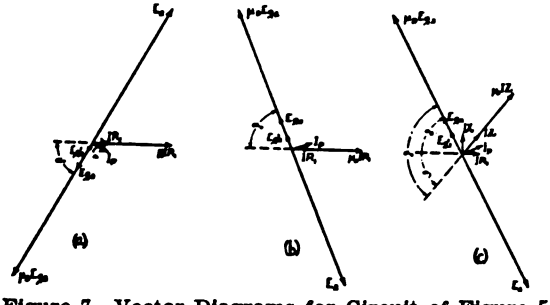

Figure 7—Vector Diagrams for Circuit of Figure 5

For a value of frequency lying between the two resonance points, the vector diagrams must be modified to those of Figs. 7 (b) and 8 (b). The equivalent transformer impedance is now capacitive; the plate current I_p therefore leads the fictitious e.m.f. $\mu_A I R_1$ and the voltage drop E_{g2} leads I_p.

As the frequency is raised above the second resonance point, it will be found impossible to effect a balance with the circuits as shown in Figs. 5 and 6, due to the fact that the phase shift angle becomes greater than 90 deg. A slight modification of these circuits, however, consisting of inserting an inductance L_1 in series with R_1 overcomes this difficulty, as may be seen from the vector diagrams of Figs. 7 (c) and 8 (c). This change in the circuit arrangements results in a corresponding change in the expressions for stage amplification and phase shift. For the circuit of Fig. 5

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{\mu_B \sqrt{R_1^2 + (wL_1)^2}} \qquad (4)$$

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$$a_0 = tan^{-1} \frac{wL_1}{R_1} + tan^{-1} \frac{wM}{R_2} \quad (5)$$

while for the circuit of Fig. 6

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{\sqrt{R_1^2 + (wL_1)^2}} \quad (6)$$

the phase shift angle being given by equation (5).

A consideration of the vector diagrams given above indicates a possible modification of the circuits of Figs. 5 and 6 which may prove quite desirable. For all frequencies above the first resonance point, a variable capacitance standard

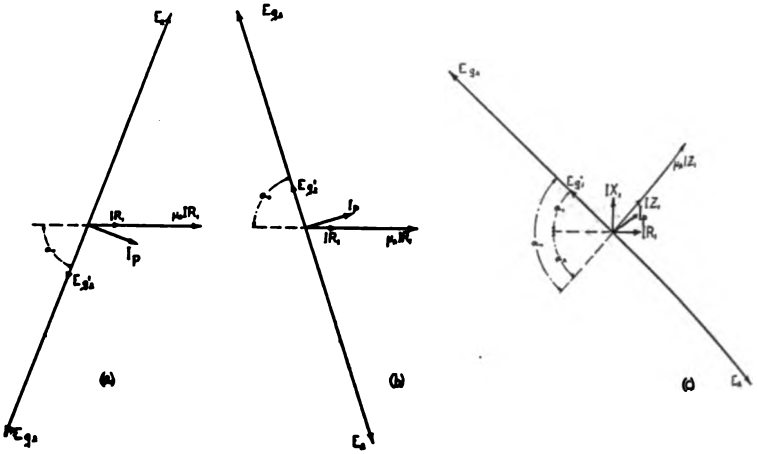

Figure 8—Vector Diagrams for Circuit of Figure 6.

connected in series with R_2 may be used to replace the mutual inductor shown. This is of particular advantage at the higher audio frequencies where errors are possible in the calibration of the mutual inductor due to its distributed capacity between turns, mutual capacity between windings, etc. Where the variable capacitance is used, a modification of the Wagner earth connection is necessary, the inductance shown being replaced by a variable condenser.

The use of the Wagner earth connection[4] is essential if any degree of accuracy is to be expected. This connection insures that the currents in the two arms R_1L_1 and R_2M do not differ from each other because of stray capacity to earth effects, an assumption upon which the expressions derived

[4]See bibliography.

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above were based. Even with the Wagner earth, however, considerable care must be taken to prevent direct capacity effects between component parts of the circuit, particularly between the two sets of batteries in the circuit of Fig. 6.

In either of the above two methods, the load on the transformer secondary due to the auxiliary tube *B* is negligible, since at the point of balance no current exists in the plate circuit of this tube. The actual loading is in each case accomplished by connecting the input circuit of a second amplifier across the transformer secondary, as shown, the

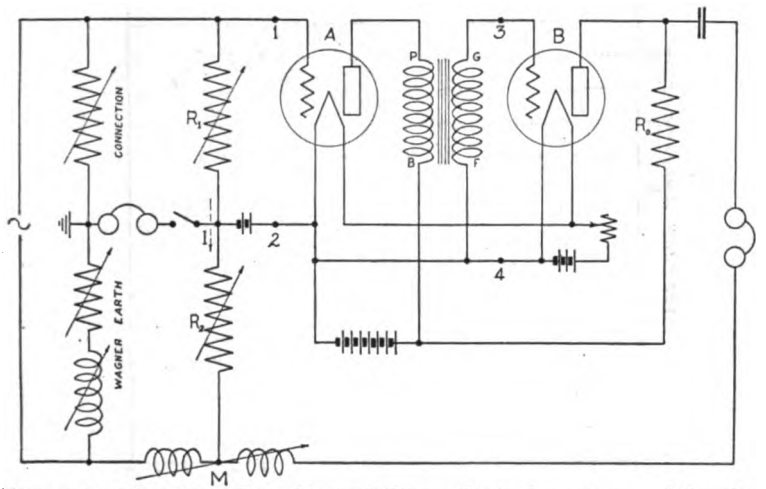

Figure 9—Simplified Test Method When Plate Impedance of Loading Tube Is Resistive.

reaction due to the impedance in the plate circuit of this second amplifier constituting the load. By varying the magnitude and character of this impedance, the effect of varying the reaction upon the amplifier performance may be observed.

For the particular case, where the effect of reaction due to a pure resistance load in the plate circuit of the tube following the transformer is desired, the circuit of Fig. 9 has been found very convenient. This method may be extended to include the reaction effect due to any impedance, whatever, but becomes somewhat involved. In its operation, it is essentially similar to the methods already described. The voltage E_2, due to IR_2 and IwM is here balanced against that part of the driving voltage in the plate circuit of tube

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B which is necessary to send the current I_{pB} through the coupling resistance R_0. The expression for the stage amplification and phase shift angle of the test amplifier is respectively

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{R_1\left\{\mu_B\left(\frac{R_0}{R_0 + r_{PB}}\right)\right\}} \tag{7}$$

$$a_0 = tan^{-1}\ \frac{wM}{R_2} \tag{8}$$

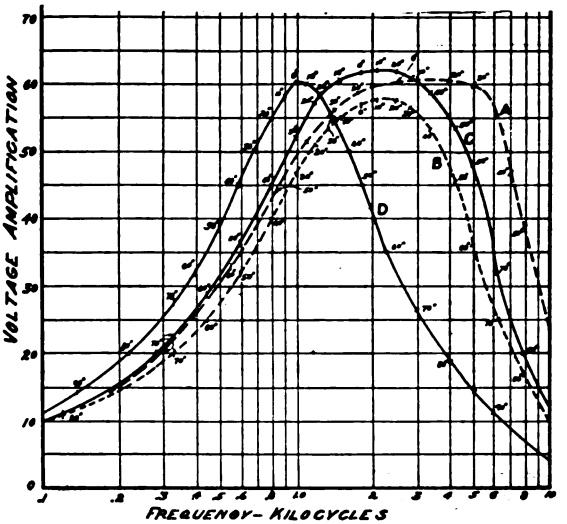

Figure 10—Effect of Load on Amplifier Performance

The value of $\mu_B \frac{R_0}{R_0 + r_{PB}}$ does not vary as long as the coupling resistance and plate battery voltage remain unchanged, and may be determined simply by disconnecting the apparatus between leads 1-2 and 3-4, connecting 1 to 3 and 2 to 4, respectively, and with M at zero value adjusting R_2 for a balance in the phones.[6] This quantity is then equal to the ratio of R_2/R_1 when the condition of balance obtains. In all other particulars, the method of Fig. 9 is exactly similar to those already described.

Results

In Fig. 10 are shown the characteristics, under various conditions of load, of a stage of transformer coupling com-

[6]See bibliography.

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prising a UX-201A tube and a transformer of rather indifferent design. Curve *A* is for the transformer unloaded, and curve *B* for a reaction load on the transformer secondary produced by a 50,000-ohm resistance connected in the plate circuit of the tube following the transformer. The input capacity C_g of the loading tube, corresponding to 50,000 ohms in its plate circuit, was measured by a resonance method and found to have a value of 85 micro-micro-farads.

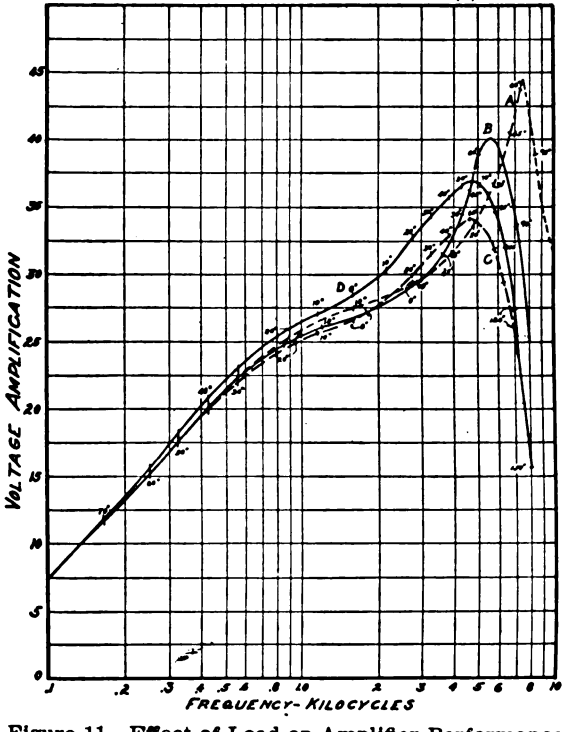

Figure 11—Effect of Load on Amplifier Performance

(This value includes capacities between connecting leads, socket terminals etc., as well as the effect of reaction). Curve *C* was then obtained for an artificial load of 85 $\mu\mu$f. across the transformer secondary. A good degree of accordance between curves *B* and *C* will be observed. The lower values of amplification obtained in *B* are no doubt due to the dielectric losses in the tube capacities rather than to the input resistance R_g due to reaction, which is too small to

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have this effect. Curve *D* corresponds to an artificial load of 250 u.u.f. across the transformer secondary.

Similar curves are shown in Fig. 11 for a transformer of somewhat different design. Maximum amplification is here seen to occur at the second resonance point. Curve *A* is for no load on the transformer, and curve *B* for reaction load due to 50,000 ohms in the plate circuit of the loading tube. For curve *C* the same resistance was inserted in the plate circuit of the loading tube and in addition a capacity of 8 m.m.f. connected between the grid and plate of this

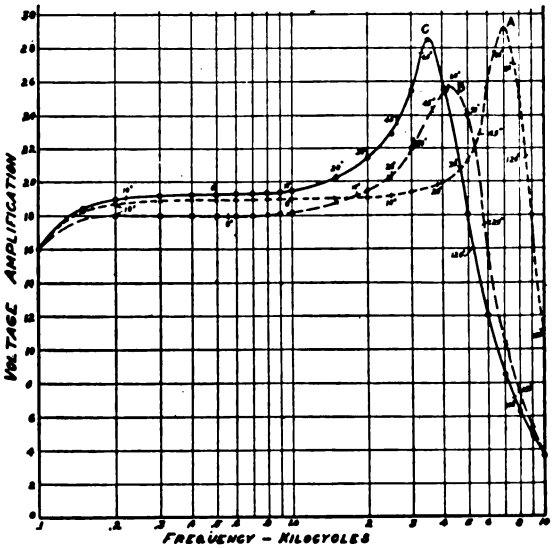

Figure 12—Effect of Load on Amplifier Performance

tube. The input capacity C_g corresponding to this condition was found to be 135 $\mu.\mu.$f. Curve *D* is for an artificial load of 135 $\mu.\mu.$f. across the transformer secondary.

An additional set of characteristics for an amplifier using one of the better types of transformers is given in Fig. 12. Curve *A* is the no-load characteristic. To obtain curve *B*, a resistance of 50,000 ohms was connected in the plate circuit of the loading tube; while for curve *C* this resistance was replaced by the transformer of Fig. 10 (with its secondary unloaded).

Referring now to Fig. 4, the impedance-frequency characteristic shown was obtained for the transformer of Fig.

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12 with an artificial load of 85 $\mu\mu$f. across its secondary. The resonance points should therefore occur at the same values of frequency as in curve *B* of Fig. 12. This is seen to be the case.

The results given above are thought sufficient to illustrate the dependence of the performance of a transformer-coupled amplifier upon the nature of the plate impedance in the tube following the transformer. In general, the no-load characteristic appears to be optimistic, particularly from the point of view of phase distortion. Furthermore, it would

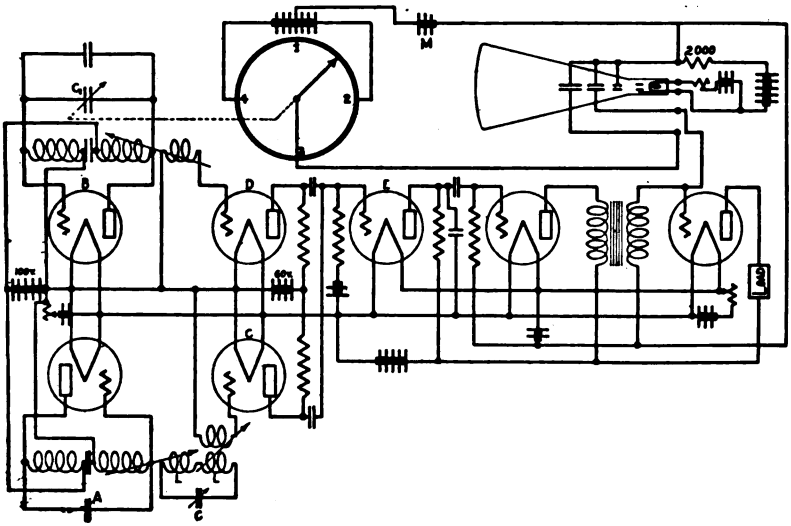

Figure 13—Schematic of Oscillographic Method of Test

seem impossible to predict from this characteristic what the performance under definite loading conditions will be. An accurate specification of the performance of a given coupling transformer in a given circuit can therefore be shown only by its actual operating characteristic as obtained above.

It is realized that there are numerous circuit conditions under which a transformer may be used and that a determination of its performance under all conditions therefore becomes impracticable. This has been used as an argument in favor of testing the amplifier with no load on the transformer secondary. As shown above, however, the no-load characteristic appears of little value.

Merely as a suggestion, the writers point out that the

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characteristic corresponding to 50,000 ohms resistance in the plate circuit of the tube following the transformer is fairly representative. Moreover, the circuit of Fig. 9 which may be used for obtaining this characteristic is very convenient and quite simple. We therefore prefer this characteristic rather than the no-load characteristic as a standard.

Section 2.—Oscillographic Method of Test

For a much more rapid determination of the amplitude-frequency characteristic, the circuit of Fig. 13 was devised:—

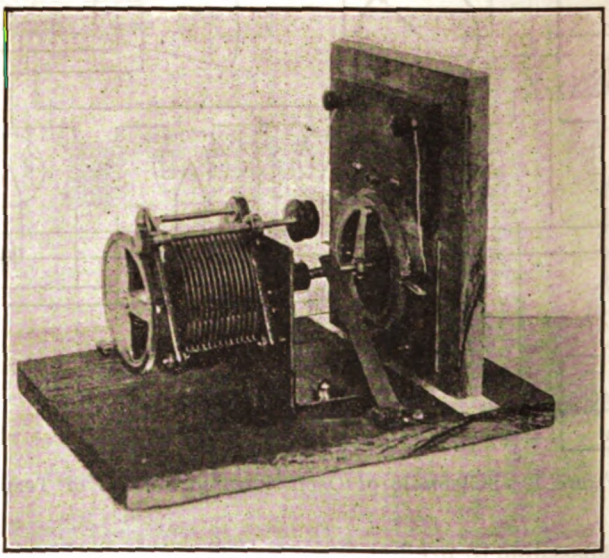

Figure 14—Illustration of Sweep Voltage Arrangement

A heterodyne oscillator is used as the source of a signal note of constant amplitude, but of a frequency varying over the important range of audio-frequencies. This signal is applied to the input circuit of the amplifier under investigation, the output voltage of this amplifier being connected across a pair of deflecting plates of a cathode-ray oscillograph. The electron beam is thus made to vibrate vertically, the amplitude of vibration being a measure of the amplitude of the output signal. Since the input to the amplifier is constant over the complete range of frequencies, the amplitude

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of vibration will vary with the frequency depending upon the amplifier characteristics.

To the other deflecting plates of the oscillograph is connected a sweep voltage arrangement whereby the electron beam is deflected horizontally, the sweep voltage being so adjusted that the spot is at the extreme left whenever the signal frequency is a minimum and at the extreme right whenever the signal frequency has a maximum value. The pattern appearing on the screen is then the desired amplitude-frequency characteristic and may be photographed or studied visually.

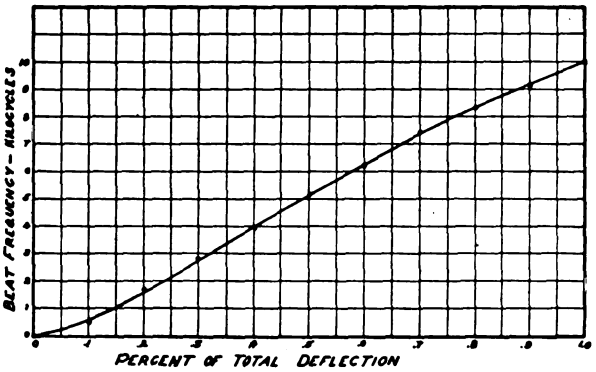

Figure 15—Frequency Calibration Curve for Heterodyne Oscillator

Just as in the test circuits of Section 1, the effect of loading the transformer secondary due to reaction may be obtained by means of an auxiliary amplifier having in its plate circuit any desired impedance.

A detailed description of the circuit of Fig. 13 is given below.

The Heterodyne Oscillator[6,7]. The two Hartley oscillators *A* and *B*, their associated amplifiers *C* and *D*, and the detector tube *E* together make up the heterodyne oscillator which is used as the variable frequency source. Oscillator *A* is adjusted to 100,000 cycles and is coupled to its amplifier *C* by means of the tuned circuit *LCL*. Oscillator *B* has a fixed coil and fixed condenser of such values, that by varying the 500 $\mu.\mu.$f. condenser C_1 from its minimum to its maximum setting, the oscillator frequency is

[6]See bibliography.
[7]See bibliography.

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reduced from 100,000 to 90,000 cycles. *B* is inductively coupled to its amplifier *D*.

The output of each of the amplifiers *C* and *D* is impressed between the grid and filament of the detector tube *E* by means of resistance-capacity coupling. The frequency of the resultant signal in the detector output circuit is, of course, the "beat" frequency of the two oscillators and should vary from 0 to 10,000 cycles as the dial of the variable condenser C_1 is tuned through an angle of 180 degrees.

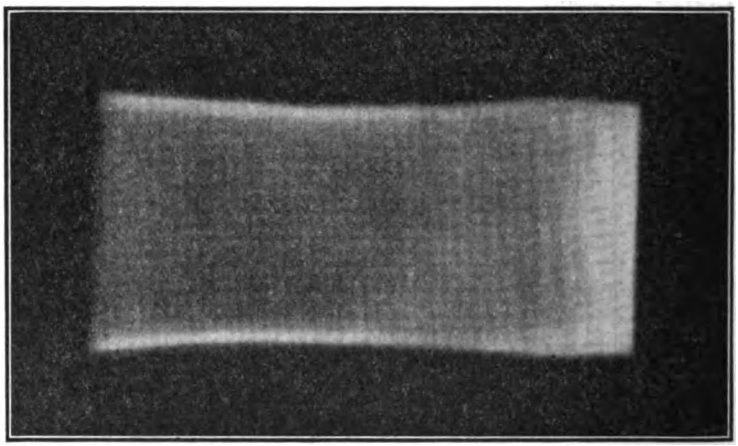

Figure 16—Amplitude-Frequency Characteristic of Heterodyne Oscillator

It is this signal of variable frequency which is impressed upon the amplifier input circuit as shown.

Actually, there is a tendency for the two oscillators to pull into step at the lower frequencies. The function of the amplifiers *C* and *D* and of the tuned circuit *LCL* is to prevent this action, a lower limit of 50 cycles per second being easily obtained.

The tuned circuit has, also, two additional advantages. Kirke has shown that in the case of two carrier waves impressed upon a non-linear detector, if the two waves are of unequal amplitude the stronger wave alone being sufficient to swing over the non-linear portion of the characteristic and reach considerably along the linear portion, the resultant detector current is proportional to the weaker wave, being unaffected by a change in amplitude of the stronger wave. By making the double coupling through the tuned cir-

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cuit quite loose and applying this loose coupling to the fixed oscillator *A*, we may make the amplitude of the fixed oscillator weaker than the weakest amplitude of the variable oscillator. A constant output is thus assured which is independent of the frequency.

Kirke has also shown that if one r.f. oscillator contains no harmonics, none will appear in the audio-frequency output except as introduced by the detector tube itself. The tuned circuit here serves to eliminate all harmonics from the fixed oscillator. The harmonics due to the detector itself

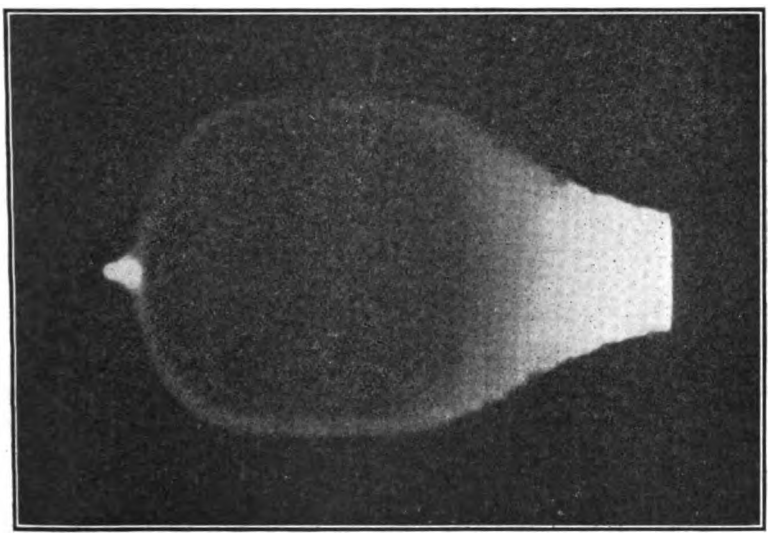

Figure 17—Oscillogram Corresponding to Curve A (Figure 10)

are minimized by using plate rather than grid rectification.

The Sweep Voltage Arrangement. As noted above, the function of the sweep voltage is to deflect the electron beam uniformly from left to right as the frequency is increased from 50 to 10,000 cycles and back to the left as the frequency is decreased to its minimum value. Furthermore it is necessary that both the change in frequency and the rate of deflection of the beam are exactly equal in order that the figure be plotted to a linear scale of abscissas, which, as may be seen from Fig. 15, is also approximately a linear scale of frequencies. Again, the spot must be in exactly the same position for a definite frequency whether the heterodyne oscillator frequency is increasing or decreasing at the instant

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considered. This is necessary in order that the pattern may exactly retrace itself.

To comply with the above conditions, the arrangement shown in Figs. 13 and 14 was devised. Mounted on the same shaft as the variable condenser C_1, but insulated from it by means of a hard-rubber insulating joint, is a movable contactor arm A making a rubbing contact with a circular carbon annulus of rectangular cross-section. This carbon annulus was constructed by placing end-to-end and joining together two circular half-rings each having a resistance between end surfaces of approximately 10,000 ohms. A 70-

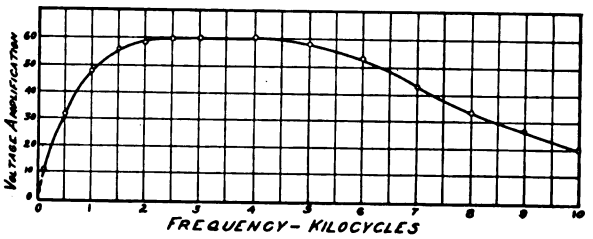

Figure 17a—Curve A Replotted.

volt battery is connected to diametrically opposite points on the circular annulus (to the joints for convenience) and two leads, one from the center in this battery and the other from the movable contactor arm, are brought out to the deflecting plates of the oscillograph. This constitutes the sweep voltage.

With the movable contactor arm at point 1, the voltage across the deflecting plates is zero. The beam should then be in the exact center of the screen, and may be so adjusted, by means of the auxiliary battery M. With the contactor arm at 2 the deflecting voltage is +35 volts, at 3 zero, and at 4 —35 volts. The electron stream is thus moved continuously from side to side, simply by revolving the contactor arm at a uniform rate.

The position of the condenser on the shaft is now so adjusted that its minimum setting occurs with the contactor at point 4 and consequently its maximum setting with the contactor at 2. The oscillator frequency thus varies from a minimum to 10,000 cycles and back as the spot travels from left to right and back.

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For an exact retracing of the figure, the capacity of the variable condenser should vary linearly over its entire range. Moreover the variable condenser should have as small a zero reading as possible. The condenser used was a General Radio Type 247-H of the geared vernier type, a groove being cut in the vernier knob which served as a driving pulley for the condenser and also the movable contactor arm. A reduction in speed was thus obtained permitting the use of a normal speed d-c. or a-c. motor.

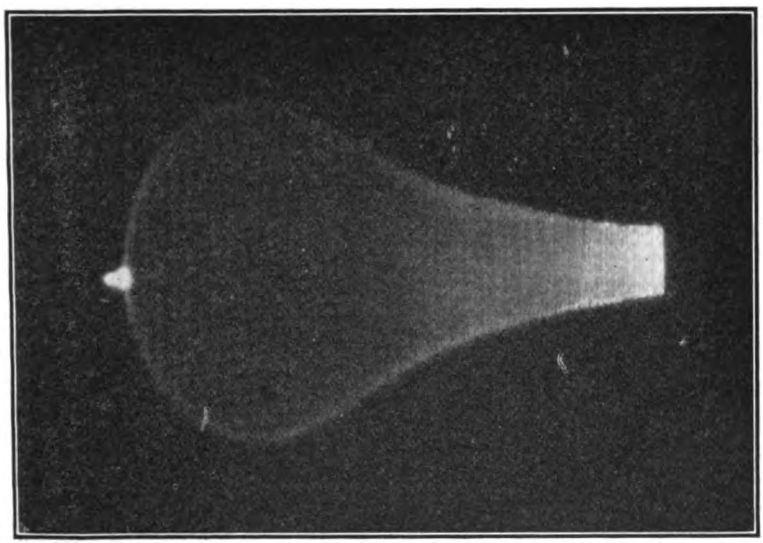

Figure 18—Oscillogram Corresponding to Curve B (Figure 10)

Oscillator Characteristics. Even though an exact retracing of the pattern on the screen is obtainable, does not imply, however, that the curve obtained is plotted to a linear scale of frequencies. The extent of deviation from a linear scale is of course of considerable importance in the analysis of any curve obtained by this method. The calibration curve of Fig. 15 plotting the beat frequency against horizontal deflection on the screen indicates however that the departure from a linear scale is fairly small.

In Fig. 16 is shown an oscillogram of the variation of output amplitude vs. frequency for the heterodyne oscillator. The amplitude is seen to be remarkably constant for the entire frequency range.

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Results

The oscillograms chosen for reproduction here are for the transformer of Fig. 10. Figs. 17, 18, and 19, correspond to curves *A*, *B*, and *D*, respectively. For the purpose of comparison these curves are replotted here to the same scale as their corresponding oscillograms. A very good agreement is observed.

Appendix

As we have already pointed out, the amplitude-frequency characteristics as obtained by the oscillographic method of test are plotted to an approximately linear scale of frequencies.

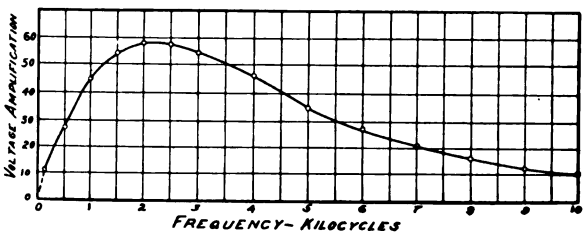

Figure 18a—Curve B Replotted

A slight modification of the sweep voltage arrangement, however, will enable the determination of these curves to a logarithmic scale of frequencies. This modification consists simply of varying one dimension of the circular carbon annulus.

A consideration of the operation of the sweep voltage device indicates that one half-ring is in operation during the time that the electron beam is being swept from left to right, the other half-ring being in use on the return sweep. We may therefore consider each half-ring alone. (see Fig. 20).

Let w be the width of the ring

t its thickness

l its length along a mean radius

and x the distance along its mean radius from the left end to the position of the contactor arm.

On the assumption of a linear characteristic for the variable condenser C_1 and a constant speed of rotation, the following equations may be set down.

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$$C_x = 500 \cdot \frac{x}{l} \tag{9}$$

Where C_x is the capacity of the variable condenser when the contactor arm is at the point x.

$$D_x = K_2 R_x \tag{10}$$

D_x being the deflection of the electron beam as measured from its extreme left position,

R_x the resistance of the half-ring from its left end to the point x and

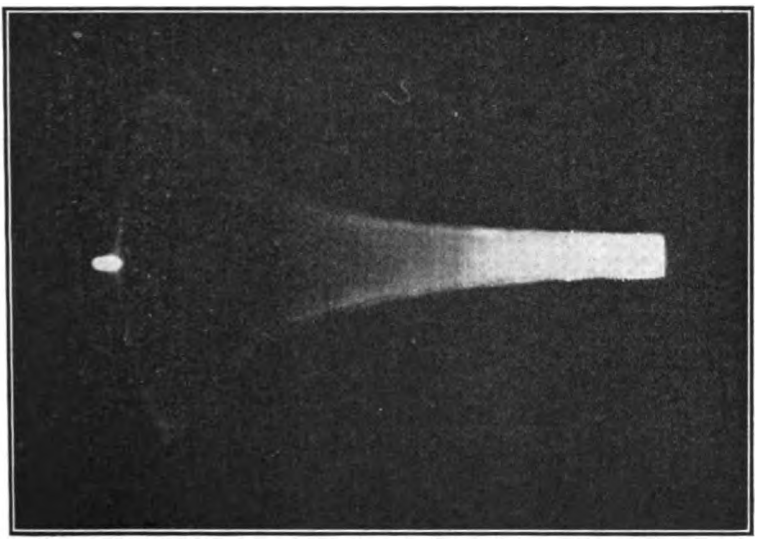

Figure 19—Oscillogram Corresponding to Curve D (Figure 10)

K_2 an arbitrary constant.

$$dR_x = \frac{\rho \cdot dx}{w \cdot t(x)} \tag{11}$$

where dR_x is an element of resistance at point x.

Note that the thickness of the ring is here taken as a function of x. It is this dimension of the ring which is to be varied in order to obtain a logarithmic scale of frequencies.

$$f_x = 100{,}000 - \frac{1}{2\pi\sqrt{K_1 L\left(2500 + 500\,\frac{x}{l}\right)}} \tag{12}$$

Here f_x is the value of the heterodyne frequency when the contactor is at the point x.

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In order that a logarithmic scale of frequencies may obtain, the deflection of the beam must vary as the logarithm of the frequency; that is,

$$D_x = K_3 \log f_x \tag{13}$$

Where K_3 is an additional constant.

Substituting equation (10) and (12) in (13) and differentiating, we obtain,

$$\frac{d\,Rx}{dx} = \frac{K}{l} \times \left\{ \frac{1.}{\left(2500 + 500 \cdot \frac{x}{l}\right)\left(1 - 2\pi \cdot 10^5 \sqrt{K_1 L}\, \sqrt{2500 + 500 \cdot \frac{x}{l}}\right)} \right\} \tag{14}$$

Where dR_x is defined by expression (11). After due substitution, equation (15) is obtained.

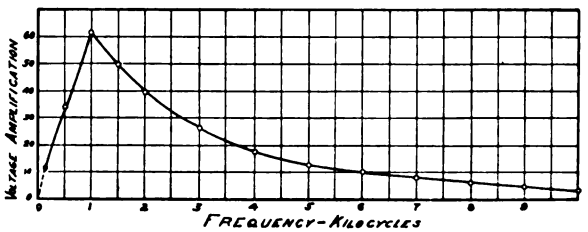

Figure 19a—Curve D Replotted

$$t(x) = \frac{e\,l}{w\,K} \left\{ \left(2500 + 500 \cdot \frac{x}{l}\right) \left(1 - 2\pi \cdot 10^5 \sqrt{K_1 L} \sqrt{2500 + 500 \cdot \frac{x}{l}}\right) \right\} \tag{15}$$

This is the necessary relation for the thickness of the carbon half-ring as a function of the distance x measured along a mean radius from the left end.

The constants of equation (15) may be evaluated by substituting for terminal conditions.

When $x = 0$

$f_x = 100{,}000$

and consequently

$$\sqrt{K_1 L} = \frac{10^{-1}}{\pi}$$

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When $x = l$

$$t(l) = \frac{e\,l}{w\,K}\left\{-(3000)\left(1-\frac{\sqrt{3000}}{50}\right)\right\} = -288$$

$$\frac{e\,l}{w\,K} = -288\,K_4$$

therefore

$$K_4 = -\frac{t(l)}{288}$$

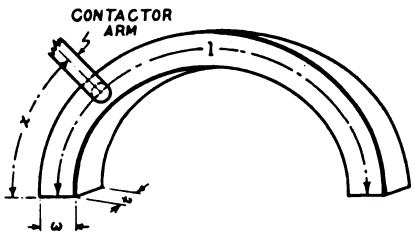

Figure 20—Carbon Half-Ring

Equation (15) may therefore be written

$$\frac{t(x)}{t(l)} = -\frac{1}{288} \times$$

$$\left\{\left(2500 + 500\,.\,\frac{x}{l}\right)\left(1 - .02\sqrt{2500 + 500\,.\,\frac{x}{l}}\right)\right\} \qquad (16)$$

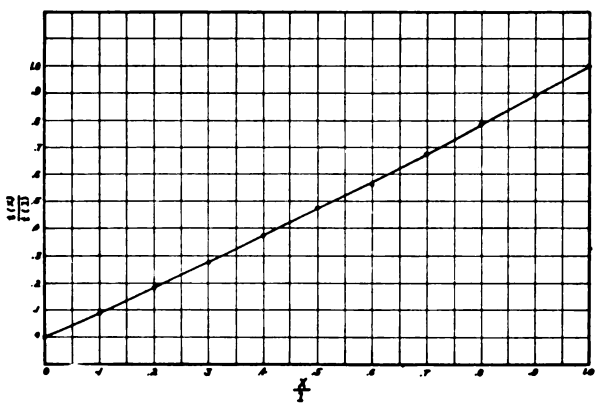

Figure 21—Computed Variation of t (x) to obtain Logarithmic Scale of Abscissas.

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This equation is plotted in Fig. 21 and is seen to be linear to a fair degree of approximation. To the same degree of approximation, therefore, we may expect eo obtain a logarithmic scale of frequency merely by tapering the thickness of the annular half-ring from zero at its left end to its normal value at the right end. This obviously applies equally as well to the other half-ring.

SUMMARY

The performance of an audio-frequency transformer-coupled amplifier is considerably affected by the reaction load across the coupling-transformer secondary due to the impedance in the plate circuit of the tube following the transformer. The effect of such reaction is discussed in this paper, and several methods of test described whereby the actual performance of a given amplifier under any condition of loading (due to reaction) may be measured.

The first two methods of test are essentially bridge circuits in which the amplifier output e.m.f. is balanced against an auxiliary voltage, the magnitude and phase relationship of which (relative to the amplifier input voltage) is determinable. To effect this balance an auxiliary tube with phones in its plate circuit is used; the balancing taking place in the plate circuit of this tube in the first method; and in its grid circuit in the second method. The reaction load across the coupling transformer is in each case obtained by connecting across its terminals the input circuit of a second amplifier, having in its plate circuit any desired impedance.

In the third method of test, the second amplifier tube serves both for loading and for balancing, the auxiliary balancing tube being omitted. This circuit is applicable to the particular case where the effect of reaction due to a pure resistance load in the plate circuit of the tube following the transformer is desired and for this case is very convenient and simple.

The fourth test method is oscillographic. A heterodyne oscillator is used as the source of a signal note of constant amplitude, but of a frequency varying over the important range of audio-frequencies. This signal is applied to the input circuit of the amplifier under investigation, the output voltage of this amplifier being connected across a pair of

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deflecting plates of a cathode-ray oscillograph. The electron beam is thus made to vibrate vertically. To the other deflecting plates of the oscillograph is connected a sweep voltage arrangement whereby the electron beam is deflected horizontally, the sweep voltage being so adjusted that the spot is at the extreme left whenever the signal frequency is a minimum, and at the extreme right when the signal frequency has a maximum value. The pattern appearing on the screen is then the desired amplitude-frequency characteristic. With a slight modification of the sweep-voltage device, this characteristic may be obtained to a logarithmic scale of frequencies. The loading is obtained in the same manner as in the methods already described.

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NOTE ON DETECTION BY GRID CONDENSER AND LEAK

By
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(Technical and Test Department, Radio Corporation of America)

Introduction

When grid condenser and leak are used for detector the grid is usually kept somewhat positive. This may make the grid-to-filament a.-c. resistance low compared with the external leak resistance. As the two resistances are effectively in parallel, both must be taken into account. A formula is derived showing the variation of detected current with frequency of modulation.

Fig. 1 shows connections commonly used for rectification by means of a grid condenser and leak. The plate circuit

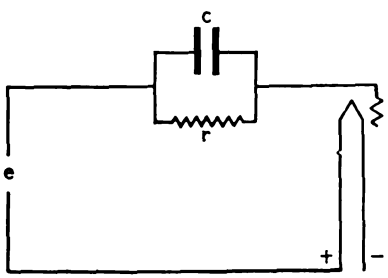

Figure 1

is not shown, as it should be so adjusted that it merely repeats the audio-frequency variation of grid potential.

Fig. 2 shows the circuit in a more general form, a battery being included to indicate that the grid return may be connected to a point of any desired potential. The impedance Z may be of any nature in the following analysis, which presupposes weak signals.

*The method used is essentially that of Carson, (Proc. of the Institute of Radio Engineers. Vol. 7, No. 2) in a paper entitled "A Theoretical Study of the Three Element Vacuum Tube."

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Let the applied voltage consist of two components, a voltage $e_0 \cos \omega_0 t$ due to the carrier, and $e_1 \cos \omega_1 t$ due to some typical side frequency. The resulting current through Z will have a number of components, but the only one that interests us is the component whose frequency is the difference between the two applied frequencies. The calculation* of this component is somewhat complicated, and for the present purpose it is sufficient to state that the magnitude of the difference frequency current varies inversely as $\left(1+\frac{Z_0}{R}\right)\left(1+\frac{Z_1}{R}\right)\left(1+\frac{Z_{1-0}}{R}\right)$ where R is the a.-c. resistance between grid and filament, Z_0 is the impedance of Z at carrier frequency, Z_1 is its impedance at the typical side frequency, and Z_{1-0} is its impedance at the difference frequency.

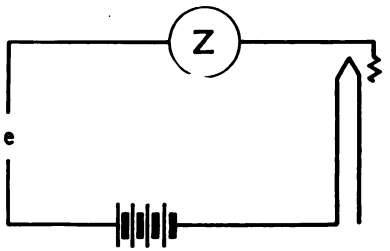

Figure 2

The voltage drop across Z due to this difference frequency current will then vary directly as $\left(\frac{R}{R+Z_0}\right)\left(\frac{R}{R+Z_1}\right)\left(\frac{RZ_{1-0}}{R+Z_{1-0}}\right)$. Now the thing we are trying to find out is how this voltage drop varies with frequency. If it is independent of frequency over the audio range then no frequency distortion results from this method of detection. Let us consider each factor separately. The first factor does not change at all, as the carrier frequency is the same no matter what the audio modulation frequency may be. The second factor is not independent of the beat frequency, but is practically so because the whole audio range may be covered by varying ω_1 about one per cent or less at broadcast frequencies. Hence unless Z contains some element sharply tuned to radio

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frequency the first two factors are practically constant whatever difference frequency is chosen. Any frequency distortion must then be due to the third factor varying with frequency. An idea that immediately suggests itself is to use for Z some network that has substantially the same impedance to all frequencies in the audio range, and low impedance to radio frequencies. However, before considering this idea further let us find out how much distortion is likely to be introduced by the third factor when Z is composed of the ordinary size grid condenser and leak, as in Fig. 1. From the form of the third factor it is obvious that it represents the impedance of R connected in parallel with Z, and hence of R, r, and C all in parallel. The impedance of this combination varies inversely as

$$\sqrt{1+\left(\frac{rR}{r+R}\ C\omega_{1\text{-}0}\right)^2}$$

As an example let us choose $C = 0.00025$ mfd. and $r = 5$ megohms, and calculate how great $\omega_{1\text{-}0}$ may be without the impedance falling below 90 per cent of the value it has at very low frequencies. If a 201A tube is used with the connections of Fig. 1 and a 5-megohm leak the value of R is about 65,000 ohms. A simple calculation shows that a frequency of about 4,500 cycles must be exceeded before the frequency distortion due to this type of detection exceeds 10 per cent. Hence such distortion is likely to be negligible in comparison with that occasioned by the use of sharply selective radio-frequency circuits. This statement has been verified by experiment, comparing the action of grid condenser and leak detection with that of C battery detection.

It is often stated that the time constant Cr should be smaller than the time of one cycle of the highest audio frequency that is to be received. If that were the case, then the combination assumed above (5 megohms and 0.00025 mfd.) would not allow satisfactory reception of audio frequencies above 800 cycles, which is contrary to experience. However the expression derived above suggests that the time constant of the condenser in parallel with the effective resistance $\frac{rR}{r+R}$ is the proper time constant to consider. Using this value and applying the time constant criterion we predict fairly good detection as high as some 60,000 cycles. This seems very high, but superheterodynes are

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often built, with similar values of grid condenser and leak for the frequency changing tube, which give a satisfactory output at intermediate frequencies of this order of magnitude.

In the foregoing it has been assumed in all the illustrative examples that the grid was kept positive. If kept negative, R is no longer low compared to r, and the proper time constant to use would be more nearly Cr. It has also been assumed that the source of applied voltage is not affected by tube losses. Actually, if the grid is kept positive, the resonance curve of the tuned circuit which supplies the input voltage will be broadened, so that less frequency distortion is produced by the tuned circuit than if C battery detection were used.

Summary

It has been said that detection by the ordinary grid-condenser and leak gives poor quality (i. e. low audio frequencies favored at the expense of the higher ones) unless an undesirably small condenser or undesirably low leak resistance is used. This paper attempts to show, as simply as possible, that actually, good quality can be obtained even when using as large a grid condenser as desirable, because the resistance which determines the relative loss of high frequencies is not the resistance of the grid leak alone, but the resistance of the grid leak and grid-filament resistance of the tube in parallel. Under usual operating conditions the latter is so low that loss of quality is very slight. The resistance of the grid leak is important chiefly in determining the grid-filament resistance of the tube, and need not be low itself.

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THE TORUSOLENOID

AN IMPROVED TYPE OF FIELDLESS COIL COMBINING THE BEST FEATURES OF THE SINGLE LAYER SOLENOID AND THE TOROID

BY

ROSS GUNN, B. S. E. E., M. S., PhD.

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During the last two or three years much good work has been done on the matter of coil design and the prediction of coil performance. This information has to do with the proper design of coil types already existing and to date there appears to have been no fundamental change in the method of winding coils for high frequency. The two general types of coils now available are the solenoids and the toroids, each type offering certain advantages and each having certain inherent disadvantages. The writer has done considerable work with coils and has devised a coil which combines all the good points of each type without incorporating any of the bad features. It is for this reason that the new coils have been called torusolenoids. Because of the novel features of the coils it is believed that information as to the construction and electrical characteristics would be of general interest to those interested in the art.

Before proceeding to the description of the coil it will be found convenient to introduce a factor which the writer has used as a direct measure of coil performance. This factor has been called the "gain" and is, in fact, simply the reciprocal of the power factor of the oscillating circuit or $\frac{2\pi fL}{R}$ where f is the frequency of the oscillations, L is the inductance and R is the equivalent series resistance. Most modern radio circuits are worked in conjunction with vacuum tubes and since vacuum tubes are controlled by potential solely, one is in general interested in the potential applied to the grid of the associated tube in terms of the e.m.f. applied to the circuit. It can easily be shown* that

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the "gain" is simply the ratio of the generated potential (i. e. the potential appearing across the coil, hence the potential applied to the grid) to the applied e.m.f. These potentials are obviously 90 deg. out of phase and their ratio as given above is always greater than unity and corresponds roughly to a circuit potential amplifying factor. Thus if a circuit has a high gain it may be considered very good whereas a low gain corresponds to a poor circuit. This idea may be easily extended to a coil if by the gain of a coil we mean the gain of a circuit made up of the coil and a condenser of negligible resistance. This factor, the gain, as defined above is also equal to the circuit selectivity or sharpness of resonance and it is therefore obvious that for broadcast reception the gain should not be excessive, not over 250. perhaps for the ordinary ear.

It is readily shown that

$$\text{Selectivity}=\text{Gain}=\frac{f_r}{f_2-f_1}$$

where f_r is the natural frequency of the oscillating circuit when in resonance with the applied frequency, f_2 is the natural frequency of the circuit *above* resonance such that the oscillating current is 70.7 per cent of the current *at* resonance and f_1 is the natural frequency of the tuned circuit at a point *below* resonance such that the oscillating current is 70.7 per cent of the current *at* resonance. Choosing therefore a mean broadcast frequency of 750 kc. and making the assumption that $G=250$; one finds that the width of the resonance curve between points where the oscillating current (and hence the generated potential) falls to 70.7 per cent of its maximum at resonance is 3,000 cycles. With $G=250$ modulation frequencies of 1500 cycles will then be amplified in potential but 70.7 per cent that of low modulation frequencies. This difference in amplification would be noticeable to a good ear but there is little doubt that the audio amplifier will partially make up this loss due to audio frequency regeneration which will accentuate the high frequencies and the arbitrary selection of 250 for the maximum value of G seems satisfactory for present day needs at least.

Returning now to the consideration of the coil, let us write down the essentials of the supreme coil. Perhaps not

*Some Facts About Coil Design, R. Gunn, Radio Broadcast, May, 1927.

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all of these can be obtained immediately but the coil to be described seems to fulfill the requirements to a surprising degree. The essentials of the ideal coil are somewhat as follows:

(1) The gain factor and hence the selectivity of the circuit as defined in the preceding paragraphs should be a maximum but should not exceed a mean value of 250 if the coil is to be used for broadcast reception and faithful reproduction desired.

(2) The external field of the coil should be zero or certainly nearly so in order to reduce to a minimum stray energy exchanges within the radio set.

(3) The distributed capacity of the coil should be very low in order that the tuning range should be a maximum and in order that the dielectric losses within the coil may be a minimum. A nearly uniform potential gradient should exist throughout the coil and the terminals should be well separated so that connecting wires will not introduce excessive capacity.

(4) The physical structure of the coils should not be excessively large taking into account the fact that a small coil necessarily means a low value for the gain under similar types of construction.

(5) Mechanically it should be strong and capable of withstanding a certain amount of abuse.

The first essential contains much implied meaning, since for the gain and the selectivity to be large, great care must be exercised in the selection of materials and so placing them as to produce no unnecessary losses. It would appear that essentials (2) and (3) are entirely incompatible and it is on this point that the great advantage of the new coils depends for it has been found possible to satisfy both conditions simultaneously by a special method of winding the coils.

The single layer solenoid which is the coil in general use today has a relatively large value for the gain when properly constructed. It has a sufficiently low distributed capacity and satisfies all requirements well enough except that relating to its external magnetic field. On this point the simple solenoid is very poor for it is well known that if a solenoid is not carefully placed at some distance from other parts of the radio set its exterior field will play havoc by causing

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undesired interaction. The external field is a double disadvantage in that the coil will easily affect other parts of the circuit and conversely other parts of the circuit easily affect it. In addition to this difficulty the field will set up eddy currents in any metal parts and dielectric losses in any insulator which will react unfavorably on the gain factor.

The toroid on the other hand has a very small exterior field but the selectivity and gain are very low. The distributed capacity is relatively large so that on the whole one is worse off by using a toroid in place of the single layer solenoid.

The points in favor and against each type of coil may be summarized as follows. The good features have been marked *plus* and the bad features *minus*.

Single Layer Solenoid.

+(a) High value for the gain and selectivity.

—(b) Large exterior field.

+(c) Moderate values for the distributed capacity.

Simple Toroid.

—(a) Very low to moderate values for the gain and selectivity.

+(b) Very small exterior field.

—(c) Large distributed capacity.

The points not mentioned are on the whole satisfactory for both types. It will be observed that one type of coil has one set of advantages and the other coil has the complementary ones, so if one could combine the advantages of each into one coil the resultant coil should be most satisfactory. The new Torusolenoids do just this; whence the name.

During the preliminary work on this subject the author saw that there was little chance of modifying the solenoid so that its exterior field would be small but there did seem to be lots of room for improvement in the toroid! The toroid was chosen as the coil that offered the best all around possibilities and much work was done in an attempt to improve it in its simple form but with scant success.

Method of Construction

The author's coil in external appearance is not unlike the usual toroid except it will be observed that the terminals of the new coil are on opposite ends of a diameter while

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in the conventional type the coil ends are adjacent. By the use of the new method the potential gradient is made nearly uniform along the coil and the high and low potential ends are well separated. This fact helps explain how the distributed capacities are kept to low values. It will be observed on closer examination that the coil is wound in two sections, the sections are wound in opposite directions, that is to say one half of the coil is a right hand spiral and

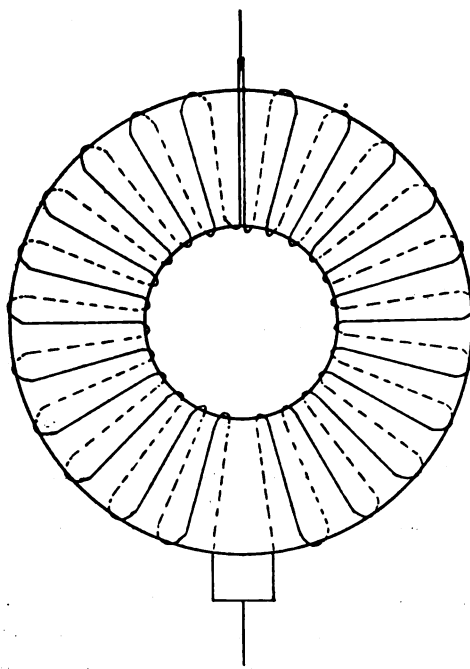

Fig. 1—Torusolenoid—The fieldless low-capacity high-gain coil.

the other half is a left hand spiral. This arrangement means that the two wires leading to either terminal *start* around the coil in the *same* direction. This particular method of winding secures a uniform magnetic field intensity within the coil so that there is no tendency whatever for the magnetic flux to stray out of the coil and cause interference with other parts of the set. The diagram of Fig. 1 shows the general method of winding much better than it is possible to describe it in words. It will be observed that each section

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or half is in itself a complete coil and the purpose of its associated coil is simply to provide a return path for the magnetic flux of the first coil. Thus each coil is mutually dependent on its mate for its efficiency. Now since the two coils are connected in parallel it is obvious that the inductance of each half must be twice the resultant inductance so that in the calculation of the number of turns necessary for a given inductance a factor of four must be introduced into the usual equation. This brings us to a third fundamental difference between the new coils and the old type. The high value of inductance for each *half* of the coil allows the use of a large number of turns of

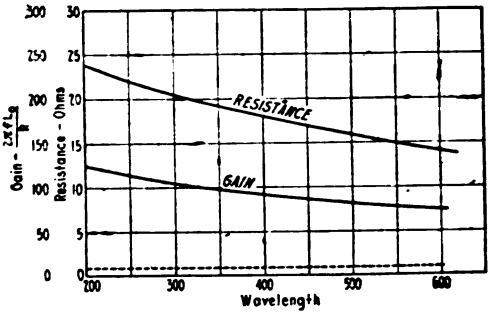

Fig. 2—Commercial "low loss" toroid
110 Turns, No. 20 B. & S.
$L_o = 330\mu h. — \lambda_o = 93.4 m. — C_o = 7.5\mu\mu f.$

relatively small wire. This is of considerable value in increasing the gain when taking into account other factors such as the fact that there are effectively two coils in parallel, for it is well known that at high frequencies the smaller wires use the copper more effectively. The use of small wire aids greatly in cutting down eddy current losses and the distributed capacity of the coil. One other feature should be mentioned and that is to point out that while the simple toroid has an exterior field corresponding to a single turn of wire whose mean area is that of the coil, the new torusolenoid has practically no field since the fields in each half tend to cancel each other outside the coil proper.

Some of the new coils have been mounted on pyrex rings and others on hard rubber forms upon which the coil and binding posts could be mounted. This method reduces to

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a minimum all losses since the coils are almost self supporting and the individual turns are properly spaced. The size of the coils have been limited to 4¼ inches since it was recognized that coils much larger than this would be unsuitable for ordinary use. By increasing the size of the coil the gain may be made much larger than the coils described in this paper and conversely if a coil is made smaller, assuming the same general type of construction the gain factor will necessarily be decreased.

The inductance may be computed for this type of coil by aid of the following relation which is similar to the usual equation save for a factor of four.

$$L_o = 0.00316\, N^2 \left[R - \sqrt{R^2 - A^2} \right]$$

where L_o is the inductance in microhenrys, R the distance from the axis of symmetry to the center of cross section of the winding, A is the radius of the turns of the winding and N is the total number of turns of the winding. All dimensions are to be inserted in centimeters. There is no known way of computing the other constants of the coil and these must be determined by experiment.

Results

Turning our attention to the actual results of tests on the improved coils we shall discuss in order the important characteristics of the coils as follows:

(a) The gain and selectivity.

(b) The exterior field.

(c) The distributed capacity.

As a basis of comparison the writer purchased a so-called low loss toroid and obtained all necessary data to rate the coil properly. It was wound with 110 turns of No. 20 B.S. wire, was self supported and had no dielectric whatever except the cotton insulation. Its direct current resistance was but 0.9. ohm while its high frequency resistance at 300 meters was 20.5. ohms. The curves of Fig. 2 show the resistance and gain for wavelengths within the broadcast band. The fundamental wavelength of the coil was 93.4 meters and its distributed capacity was 7.5 micro-microfarads.

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GAIN

It will be remembered that the gain is a function of the circuit constants only and is equal to

$$G = \frac{2\pi f L_o}{R} \quad \text{or} \quad G = \frac{1}{R}\sqrt{\frac{L}{C}}$$

where the symbols have the usual meaning. This factor may be determined directly by means of a vacuum-tube voltmeter or it may be computed from known data on the coil. The value of the gain will obviously depend on the coil and its numerical value will range from perhaps 60 for a poor coil to 300 for a high-class coil. A value much over 250 is undesirable for broadcast reception since the quality of reproduction would be impaired due to the suppression

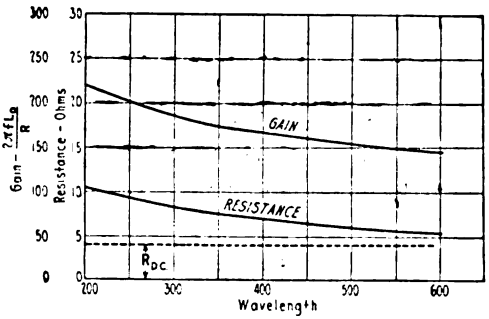

Fig. 3—Torusolenoid
480 Turns, No. 30 D. S. C.
$L_o = 242\ \mu h.$
$\lambda_o = 63 m.$
$C_o = 4.6\ \mu\mu f.$

of the side bands. The new torusolenoids have values for the gain from 135 to 250 for solid wire-wound coils and as high as 300 for litzendraht coils of moderate size. The curves of Fig. 3 show how the gain and effective series resistance of a typical solid wire-wound torusolenoid changes with wavelength. It is interesting to note that the average high frequency resistance of the torusolenoids is only about twice its direct current resistance whereas the sample of the simple toroid has a high frequency resistance about 18 times the d.-c. resistance. The data for a litzendraht coil which is typical of the class are shown in Fig. 4 and are plotted in solid lines. It will be seen that the gain over the broadcast

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band averages well over 200 and rises to well over 250 at 550 meters. The dotted curves of Fig. 4 show plainly the effect of enclosing the coil in a metal can and is remarkable in that the gain is decreased by only about 18 per cent and rather definitely showing that the exterior magnetic field is zero.

An examination of the curves, which have all been plotted to the same scale, shows that the solid wire-wound torusolenoids have the highest gain at the short wavelengths and decrease as the wavelength is increased. On the other hand the litzendraht coils show a relatively lower value at the short wavelengths and increase uniformly with increasing wavelength. This type of characteristic seems to offer

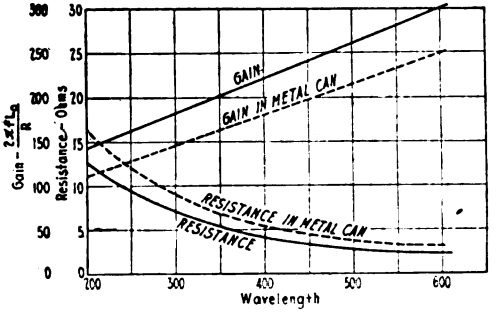

Fig. 4—Torusolenoid
300 Turns—"Litz."
In Air $L_0 = 194\mu h.$
Solid Curves $\lambda_0 = 52m.$, $C_0 = 4\mu\mu f.$
In Metal Can $L_0 = 194\mu h.$
Dotted Curves $\lambda_0 = 81.3m.$, $C_0 = 9.6.$

certain advantages since in general the energy transferred to a coil is greater at high frequencies with the usual methods of coupling, so a lower value for the gain can be tolerated. By proper design it would appear that uniform amplification could be obtained over the entire broadcast range since the gain seems to be nearly a linear function of the wavelength and the induced e.m.f. a linear function of the frequency. More work is being done on this matter in order to verify this prediction since a straightforward electrical solution of this problem is more desirable than mechanical means of varying the coupling.

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The peculiar characteristic just described together with gain factors at least equal to the simple solenoid seems to indicate that the new coils are at least on the par with any other type when considering their gain.

External Field

The external field of the new coil is even less than that of the ordinary toroid since the ordinary toroid has a field associated with it which is equivalent to a single turn whose area is equal to the mean area of the coil. In the torusolenoid the current divides and the external field of each half tends to neutralize the other.

In order to test experimentally whether the torusolenoid had an external field, a test coil was properly connected to a small fixed condenser and a sensitive thermogalvanometer. This combination was tuned to resonance with a powerful oscillator. The coil was then placed directly in the field of an oscillator plate coil which was carrying 10 amperes of a frequency corresponding to the tuned circuit of the coil. The coil under test was oriented in all directions but no current could be detected in the coil circuit. This test appears to show that the external field of the torusolenoid may be said to be effectively zero and hence stray energy exchanges due to magnetic effects are nonexistent.

Distributed Capacity

It has been recognized for a long time that one of the essentials of a good coil, regardless of its type, was a low distributed capacity. A low distributed capacity guarantees that the electric gradient within the coil has been kept small and the size and distribution of the wire used in winding has been well chosen. A low distributed capacity is particularly valuable these days when many broadcast stations are at the extremes of the band and the band none too narrow. The torusolenoids have shown an exceptionally low value for the distributed capacity, so low in fact that the usual method of plotting λ^2 against the capacity and extrapolating could not be used. The procedure used was to determine the fundamental wavelength of the coil by a simple means and then compute the distributed capacity from the relation

$$C_d = \frac{\lambda_o^2}{3.55\,L}$$

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where C_d is the distributed capacity in micro-micro-farads, λ_o is the fundamental wavelength of the coil and L is the inductance in microhenrys.

The average torusolenoid will show distributed capacities of from 4 to 5 micro-micro-farads when properly constructed. The data on the commercial coil are included in Fig. 2 for comparison.

There are many variations which may make use of the new type winding. One method which has been found fairly good is to make use of two slender solenoids which were wound in opposite directions and connected in parallel. This type of mounting yields a coil whose characteristics are

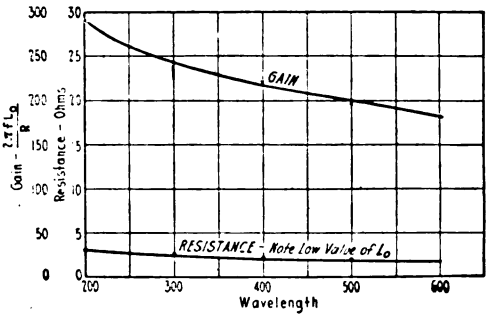

Fig. 5—Binocular—New Type Winding
No. 26 D.C.C. on 2 inch form—4 inches long.
$L_o = 93\mu h$.

shown in Fig. 5. It will be observed that the gain may be made quite high but here again there is an appreciable exterior field and a choice made as to what characteristics are most desirable.

Consideration of all the factors involved in these coils and their properties show that the torusolenoid fulfills to a remarkable degree the requirements of the ideal coil. The features possessed by the torusolenoids may be summarized as follows:

(a) They have a high value for the gain and selectivity.

(b) Their external magnetic field is vanishingly small.

(c) The distributed capacity is remarkably low.

(d) They are moderate in size taking into account item (a).

(e) They are mechanically strong enough to withstand a reasonable amount of abuse.

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SUMMARY

An improved high frequency inductance of radical design is described which represents the successful attempt to combine into one inductance all the desirable features of both the toroid and the single layer solenoid.

The new inductance which has been called the "Torusolenoid" has been found to incorporate the following features:

1. It has substantially zero external magnetic field.
2. It has a high value for the "gain".
3. It has very low distributed capacity.
4. It has no major disadvantages when used as an inductance.

The term "gain" is defined and shown to be a measure of coil performance. Supporting data on representative inductances incorporated in the paper show that the new inductances offer many advantages, and for all-around performance are unsurpassed.

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BOOK REVIEW

Robison's Manual of Radio Telegraphy and Telephony

By

COMMANDER (now REAR ADMIRAL) S. S. ROBISON,
U. S. Navy.

Revised by COMMANDER S. C. HOOPER, U. S. Navy and
LIEUTENANT COMMANDER T. A. M. CRAVEN, U. S. Navy.

Seventh Revised Edition; 737 pp.; 424 fig.; The United States Naval Institute, Annapolis, Md.; Price $5.50.

This book is divided into four sections headed in order, (1) theory of radio communication, 462 pp.; (2) application, 125 pp.; (3) radio measurements and precision instruments, 92 pp.; (4) useful information, 46 pp.

Section 1 dealing with the theory of radio communication includes the elementary theory of electricity, alternating current theory, application of alternating current theory to radio circuits, theory of damped oscillations, theory of vacuum tubes, theory of wave propagation, and radio accessories such as inductance coils, condensers and quartz crystals. A chapter on quartz crystals and one on the Taylor-Hulburt theory of transmission appear as additions in this revision. This section presents the principles underlying radio communication in a clear and concise manner.

Section II, presenting the applications, includes radio transmission, radio reception, and radio compass. This section deals with descriptions of apparatus for transmis-

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sion, reception, and radio compass. The work on spark and arc transmitters is abbreviated in this edition. The work on receivers is largely rewritten.

Section III on radio measurements and precision instruments consists of twenty-six experiments of practical and precise measurements. These are suitable for teachers in need of such material as well as for laboratory workers. This section is almost identical with the corresponding work in the 6th edition.

Section IV, useful information, includes tables and formulas which are needed by the radio student. In this edition this section has been shortened by the omission of the work on elementary mathematics, and on laws and regulations.

Taking the book as a whole it is very readable. The explanations are remarkably clear and concise. Figures are good, well placed, and used to advantage in connection with the text. The many worked examples are helpful. Mathematics is not avoided but is used in such a way that the inexperienced reader can understand it easily. Important statements are emphasized by bold-faced type. The binding is sturdy. The Manual may be highly recommended not only to the elementary student to whom it is addressed chiefly, but also to the more mature worker and teacher.

The following errors in the statements of formulas and mathematical work should be noted:— p. 35 $R = lSP$ should be $R = \frac{lP}{S}$; p. 36 $g = \frac{\gamma}{lS}$ should be $g = \frac{\gamma S}{l}$, and $g = \frac{1}{lSP}$ should be $g = \frac{S}{lP}$; p. 93 $blB_\epsilon \cos \omega (t + dt) = blB_\epsilon \cos \omega t \sin \omega dt - \sin \omega t \cos \omega dt$ should be $blB_\epsilon \cos \omega (t + dt) = blB_\epsilon (\cos \omega t \cos \omega dt - \sin \omega t \sin \omega dt)$; p. 167 line beginning "and substituting", 10^2 should be 10^4 p. 176 $I = .701\ I_0$ should be $I = .707\ I_0$; p. 225 $\frac{R}{2L} < \frac{1}{LC}$

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should be $\frac{R}{2L} < \frac{1}{\sqrt{LC}}$; p. 642 $\lambda_1 = 1885 \sqrt{(L_1 - L)C}$ should be $\lambda_1 = 1885 \sqrt{(L_1 + L)C}$; and $\lambda_2 = 1885 \sqrt{(L_2 - L)C}$ should be $\lambda_2 = 1885 \sqrt{(L_2 + L)C}$.

S. S. Kirby.

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DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued August 2, 1927—August 23, 1927

By
JOHN B. BRADY

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,637,445—A. W. FRANKLIN, New York, N. Y. VARIABLE MICA CONDENSER. Filed Oct. 11, 1923, issued Aug. 2, 1927. Assigned to Chas. Freshman Co., Inc.

1,637,613—F. A. KOLSTER, Palo Alto, Calif. RADIO COMPASS. Filed Nov. 22, 1920, issued Aug. 2, 1927. Assigned to Federal Telegraph Co.

1,637,646—H. KIMURA, Los Angeles, Calif. Filed Jan. 7, 1925, issued Aug. 2, 1927.

1,637,864—F. A. KOLSTER, Palo Alto, Calif. ELECTRON TUBE APPARATUS. Filed May 25, 1926, issued Aug. 2, 1927. Assigned to Federal Telegraph Co.

1,637,615—F. A. KOLSTER, Palo Alto, Calif. RADIO COMPASS. Filed Nov. 22, 1920, issued Aug. 2, 1927. Assigned to Federal Telegraph Co.

1,638,158—ELECTRIC APPARATUS. P. C. HEWITT, Ringwood Manor, N. J. Filed July 30, 1920, issued Aug. 9, 1927. Assigned to Cooper Hewitt Electric Co.

1,638,278—CONTACT RECTIFYING DEVICE. W. O. SNELLING, Allentown, Pa. Filed Feb. 8, 1923, issued Aug. 9, 1927.

1,638,417—SHIP'S SIGNALING OR BROADCASTING DEVICE. E. A. SPERRY, Brooklyn, N. Y. Filed Nov. 6, 1923, issued Aug. 9, 1927. Assigned to Sperry Gyroscope Company.

1,638,608—WAVE FILTER RECEIVING CIRCUIT. PETER WELCH and AUDLEY MURPHY, of Olean, N. Y. Filed Feb. 6, 1925, issued Aug. 9, 1927.

1,638,320—SOUND REPRODUCING APPARATUS. CYRIL A. BRIGHAM, of East Orange, N. J. Filed Nov. 2, 1926, issued Aug. 9, 1927. Assigned to Brandes Laboratories, Inc.

1,638,499—ELECTRON DISCHARGE DEVICE. A. MAVROGENIS, Milwaukee, Wis. Filed Jan. 2, 1926, issued Aug. 9, 1927.

1,638,699—A. L. MOBURG, Palms, Calif. UNDERGROUND ANTENNA. Filed Mar. 30, 1926, issued Aug. 9, 1927.

1,638,551—ELECTRON DISCHARGE DEVICE. V. L. RONCI, Brooklyn, N. Y. Filed July 30, 1924, issued Aug. 9, 1927. Assigned to Western Electric Co. Inc.

1,638,925—HIGH FREQUENCY SIGNALING SYSTEM. L. ESPENSCHIED, Hollis, N. Y. Filed Aug. 16, 1922, issued Aug. 16, 1927. Assigned to American Telephone and Telegraph Co.

1,638,930—VOLTAGE REGULATOR FOR VACUUM TUBES. J. HERMAN, New York City. Filed Nov. 26, 1923, issued Aug. 16, 1927. Assigned to American Telephone and Telegraph Co.

1,639,414—RADIO RECEIVING SYSTEM. J. M. MILLER, Philadelphia, Pa. Filed June 23, 1926, issued Aug. 16, 1927. Assigned to Atwater Kent Mfg. Co.

1,639,597—CONDENSER. WILLIAM DUBILIER, New York, N. Y. Filed May 23, 1922, issued Aug. 16, 1927. Assigned to Dubilier Condenser Corp.

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1,639,650—HIGH POTENTIAL CONDENSER. WILLIAM DUBILIER, of New York, N. Y. Filed March 7, 1919, issued Aug. 23, 1927. Assigned to Dubilier Condenser Corp.

1,639,667—METHOD FOR RADIO POSITION FINDING. R. H. RANGER, of Brooklyn, N. Y. Filed March 8, 1924, issued Aug. 23, 1927. Assigned to Radio Corp. of America.

1,639,686—WIRELESS TELEGRAPH STATION. H. J. J. M. DeBELLESCIZE, of Toulon, France. Filed August 29, 1921, issued Aug. 23, 1927.

1,639,695—ARRANGEMENT FOR THE REGULATION OF THE FREQUENCY IN FREQUENCY CHANGERS. K. HEEGNER, of Berlin, Germany. Filed Oct. 23, 1925, issued Aug. 23, 1927. Assigned to Gesellschaft Fur Drahtlose Telegraphie M. B. H.

1,639,696—MEANS FOR CHANGING WAVELENGTHS. G. HILL, of Brooklyn, N. Y. and GEORGE H. CLARK, of Washington, D. C. Filed May 3, 1915, issued Aug. 23, 1927. Assigned to Radio Corporation of America.

1,639,698—ELECTRON EMITTING CATHODE AND PROCESS OF PREPARING SAME. F. HOLBORN, of Hoboken, N. J. Filed Jan. 14, 1926, issued Aug. 23, 1927. Assigned to Hazeltine Corporation.

1,639,727—AERIAL. M. E. FRASER and LOUIS SCHULTZ, of Detroit, Mich. Filed Feb. 23, 1926, issued Aug. 23, 1927.

1,639,773—TWO-WAY TELEPHONE TRANSMISSION. H. S. HAMILTON, of New York, N. Y. Filed Nov. 24, 1926, issued Aug. 23, 1927. Assigned to American Telephone and Telegraph Co.

1,639,805—RADIO APPARATUS. F. S. McCULLOUGH, of Cleveland, Ohio. Filed Jan. 9, 1920, issued Aug. 23, 1927. Assigned to Glenn L. Martin.

1,639,816—RADIO SIGNALING CIRCUIT. A. H. TAYLOR and LEO C. YOUNG, of Washington, D. C. Filed Jan. 7, 1925, issued Aug. 23, 1927. Assigned to Wired Radio, Inc.

1,639,817—PIEZO ELECTRIC CRYSTAL SYSTEM. A. H. TAYLOR, of Washington, D. C. Filed July 29, 1925, issued Aug. 23, 1927. Assigned to Wired Radio, Inc.

1,639,913—ANTISTATIC AERIAL. R. A. WEAGANT, of New York, N. Y. Filed Oct. 6, 1920, issued Aug. 23, 1927. Assigned to Radio Corporation of America.

1,640,200—MOUNTING MEANS FOR RADIO RECEIVING SETS. F. L. LORD, of Newark, N. J. Filed Aug. 28, 1925, issued Aug. 23, 1927. Assigned to The Lord Laboratories, Inc.

1,640,234—ANTENNA FOR USE WITH RADIO SETS. M. M. CONYERS, of Custer, South Dakota. Filed June 14, 1926, issued Aug. 23, 1927.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED AUGUST 4, 1927

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	Allentown, 1123 Hamilton St.	Woodling, C. E.
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	Connellsville, 302 S. Pgh. St.	Angle, B. C.
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	Pottstown, 524 Charlotte St.	Lessig, L. G.
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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 15 OCTOBER, 1927 Number 10

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INSTITUTE ACTIVITIES

Meeting of Board of Direction

At the meeting of the Board of Direction held in the offices of the Institute on September 7, 1927, the following were present: Ralph Bown, President; Melville Eastham, L. A. Hazeltine, R. A. Heising, J. V. L. Hogan and J. M. Clayton, Assistant Secretary.

The Board approved the action of the Committee on Admissions in the case of the following applications: transfer to the grade of Fellow, Pendleton E. Lehde and H. E. Hallborg; transfer to the grade of Member, Ernest V. Amy, C. C. Harris, Harold Herbert, Ross A. Hull, and W. A. Thomas; election to the grade of Member, W. H. Bailey, T. W. Bearup, R. B. Owens, and John Murchie.

Eighty-four Associate and eighteen Junior members were elected.

During the months of July and August two hundred and thirty-seven Associate and thirty-nine Junior members were elected.

The Board approved the application from members residing within the vicinity of Buffalo for the formation of a Buffalo-Niagara Section of the Institute.

New York Meeting of the Institute

At the first Fall New York meeting of the Institute, held on September 7th in the Engineering Societies Building, 33 West 39th Street, New York, a paper by Messrs. H. Diamond and J. S. Webb was presented by Mr. Diamond. The subject was, "The Testing of Audio-Frequency Transformer-Coupled Amplifiers."

The following took part in the discussion which followed the reading of the paper: Professor L. A. Hazeltine, R. R. Batcher, George Crom, I. G. Maloff, Melville Eastham, and others.

The attendance at this meeting was over three hundred.

News of the Sections

Practically all of the Sections are making plans for resumption of activities for the Fall season. All of the Sections have meetings planned for the month of September.

Those Sections requiring them are now being supplied with preprint copies of all papers which are to be presented before the New York meetings. In most cases these papers will be available

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for Section use several weeks prior to the New York meetings. Sufficient preprints are being supplied for each person attending each Section meeting to secure one at the meeting.

Committee Work

At the meeting of the Committee on Admissions held on the afternoon of September 6, 1927 in the offices of the Institute, the following were present: Professor L. A. Hazeltine (Acting Chairman) and Messrs. R. A. Heising and H. F. Dart.

The Committee acted upon thirty-five applications for transfer or election to the grades of Fellow and Member.

Fourteen Year Index

Printing of the fourteen year Index to the PROCEEDINGS has been held up unavoidably. It is hoped that this Index will be in the mails shortly after the October issue is published. When completed, each member of the Institute will be mailed a copy of the Index.

The price to non-members will be one dollar.

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LONG-WAVE RADIO MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1926, WITH SOME COMPARISONS OF SOLAR ACTIVITY AND RADIO PHENOMENA*

By

L. W. Austin

(Laboratory for Special Radio Transmission Research conducted jointly by the Bureau of Standards and the American Section of the International Union of Scientific Telegraphy).

The following is a résumé of the measurements made by the Bureau of Standards on long-wave signal intensities and atmospheric disturbances during 1926, to which have been added some

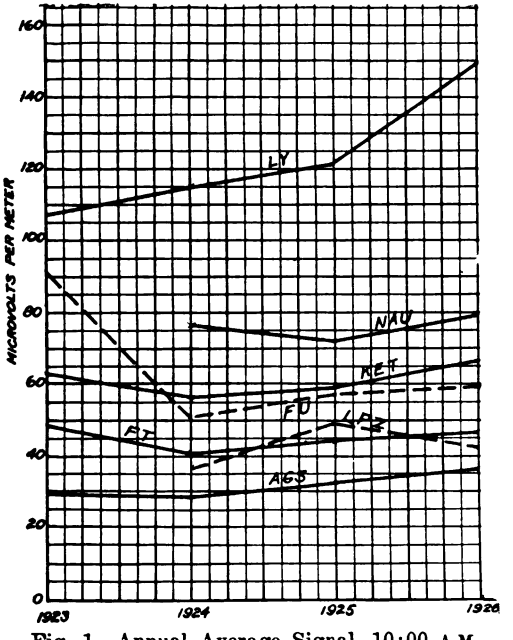

Fig. 1—Annual Average Signal, 10:00 A.M.

measurements from former years for the purpose of studying the relations of radio transmission and atmospheric disturbances to other natural phenomena.

Received by the Editor, July 12, 1927.

Read before the International Union of Scientific Radio Telegraphy, American Section, April 21, 1927.

* Published by Permission of the Director of the National Bureau of Standards of the U. S. Department of Commerce.

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The method of measuring weak signals through heavy atmospherics, described in the report for 1925[1] in which a correction factor for the deadening effect of the atmospherics is determined by measuring the apparent strength of a correspondingly weak artificial signal, with and without atmospherics, has proved entirely successful.

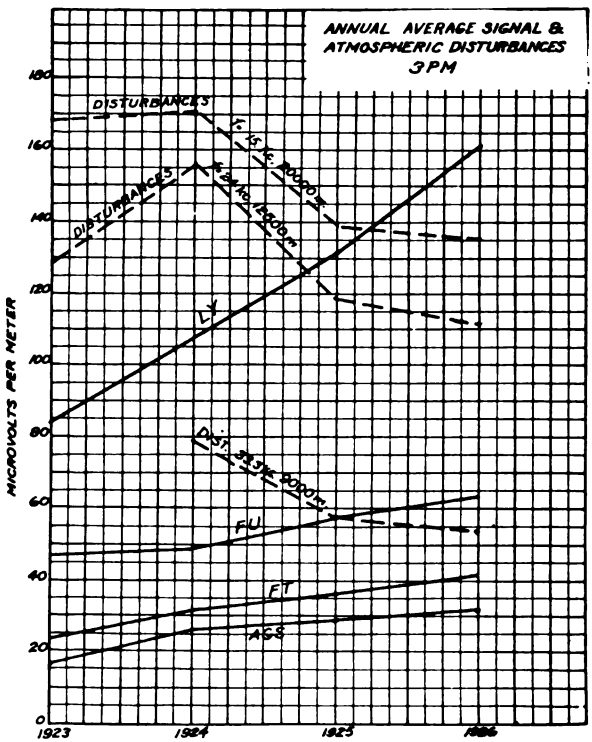

Fig. 2

On account of the lack of assistants in the laboratory for taking night and holiday observations, much time has been devoted by my assistant, Mr. Judson, to experiments on automatic measuring apparatus. At the close of the year these experiments had so far progressed that regular 24-hour records were being made of the Bordeaux, Tuckerton, and Cape Cod signals and atmospheric disturbances, using a high antenna for Bordeaux and a lower one for the American stations. A Cambridge-Paul recorder is being used which records at every half minute for five minutes in each

[1] Proc. I. R. E., vol. 14, p. 663; 1926.

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hour on each station being measured, the antenna and secondary circuits being changed in tune by means of a clock-controlled relay and secondary relays. The capacity of the system will permit further stations to be added when it seems desirable. It is, of course, not expected that European stations can be recorded during the summer without interruption from atmospherics. Apparatus is also being developed for recording the direction of atmospherics.

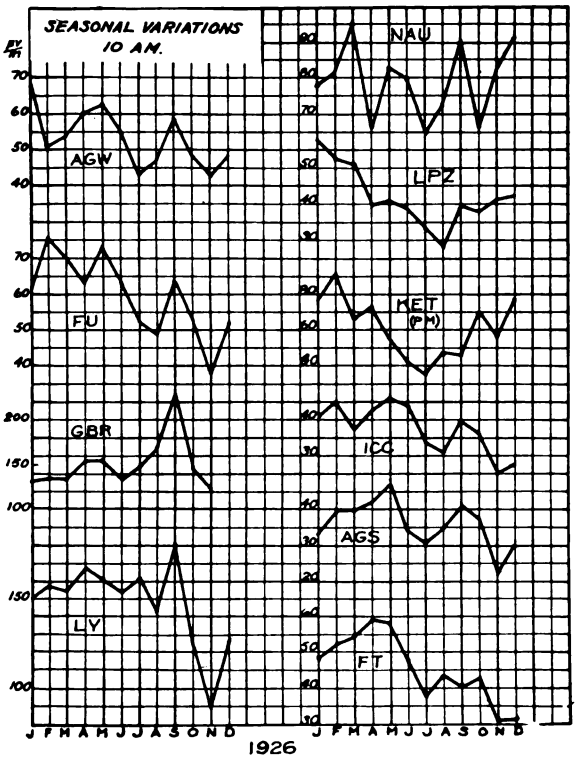

Fig. 3

The tables and Figs. 1 to 4 are similar to those given in former reports and are in general self-explanatory. The daylight strength of all the stations shown, except LPZ, has continued to rise during 1926 while the average daylight atmospheric disturbances have continued to fall.

Fig. 3 shows the seasonal variations of the 10 A.M. (E.S.T.) all daylight signals of various stations. Practically all these stations show a peak in the curve of monthly averages in Septem-

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ber or October, while the European stations show low values in November and generally in December and January. A number of stations also have peaks at some time in the spring. The north-south stations NAU (Porto Rico) and LPZ (Argentina) do not show the late autumn and winter minima nor does KET (Bolinas, Calif.) (3 P.M.)[2] although its transmission is west-east. This confirms our earlier surmise that the low autumn and winter signals of the European stations are due to the proximity of the European sunset to the transmitting time of the signals received at 10 A.M. The September or October peak may possibly have some significance. It is conceivable that it may have some connection with the autumn peak of the curve of terrestrial magnetic range.

TABLE I

Average Signal Intensity and Atmospheric Disturbances for Lafayette (LY), Ste. Assise (FU) Nauen (AGW), Rugby (GBR), Monte Grande (LPV) and Rio de Janeiro (SPR), in microvolts per meter.

	A. M.							P. M.						
1926	LY	FU	AGW	GBR	LPV	SPR	Dist.	LY	FU	AGW	GBR	LPV	SPR	Dist.
January	150	62	68	134	—	—	34	240	91	116	—	—	—	48
February	157	76	51	136	—	—	32	186	65	72	170	—	—	54
March	156	70	54	137	—	—	38	203	83	63	159	—	—	59
April	168	63	60	156	—	—	65	174	73	53	129	—	—	96
May	162	74	62	155	51	—	41	144	59	46	127	31	—	148
June	154	63	56	134	42	—	40	109	45	40	94	25	—	176
July	161	52	44	145	31	40	40	113	42	32	110	—	—	232
August	142	49	47	165	27	36	44	96	40	39	107	15	30	435
September	181	64	59	230	38	38	34	142	47	42	181	—	18	209
October	124	54	49	145	33	36	39	122	58	53	152	34	27	82
November	90	37	43	122	29	37	39	148	58	55	149	27	30	54
December	129	52	49	—	34	43	27	256	110	89	310	—	37	37
Average	148	60	54	151	36	38	39	161	64	58	154	26	28	136

Last year much time was devoted to a study of the observational data of the laboratory in regard to a relationship with temperature and this relationship was quite definitely established as far as stations at moderate distances are concerned.[3]

At present the relationships of radio phenomena to solar activity and terrestrial magnetism are being examined. Some preliminary results of this study are here given.

Solar Activity and Radio Phenomena

It has been often suspected that a connection exists between radio and the aurora and magnetic storms. The connection with magnetic storms was first definitely established by Espenschied, Anderson and Bailey[4] in the work of the Bell Telephone Company in preparation for the establishment of a transatlantic telephone

[2] The 3 P.M. signals from Bolinas are given instead of those at 10 A.M. as these give better all daylight conditions.

[3] Proc. I. R. E., vol. 14, p. 781; 1926.

[4] Proc. I. R. E., vol. 14, p. 7; 1926.

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service. They found that magnetic storms greatly decreased the strength of night signals and slightly increased the daylight strength. This effect was more pronounced at a wave length of 5000 m. than at 17000 m. Since terrestrial magnetism is known

TABLE II

Average Signal Intensity and Atmospheric Disturbances for Ste. Assise (FT), Bolinas (KET, Nauen (AGS), Monte Grande (LPZ), Leafield (GBL) and Coltano (ICC) in microvolts per meter.

	A. M.							P. M.						
1926	FT	KET	AGS	LPZ	GBL	ICC	Dist.	FT	KET	AGS	LPZ	GBL	ICC	Dist.
January	48	72	33	57	20	41	27	54	76	47	43	22	—	39
February	53	70	40	53	31	45	27	54	91	41	36	20	—	44
March	55	69	40	51	25	37	33	56	68	40	31	19	—	47
April	60	71	42	40	27	43	57	49	71	34	33	22	24	84
May	59	76	47	41	23	46	33	38	55	27	17	15	31	129
June	48	62	35	39	21	44	35	21	41	21	15	12	22	153
July	39	57	31	34	16	34	34	25	36	17	17	15	21	192
August	44	—	35	28	14	31	42	27	48	22	28	—	23	369
September	42	58	42	40	20	40	27	26	47	—	—	14	24	128
October	43	67	37	38	24	36	34	39	70	31	27	21	29	72
November	32	63	23	41	15	25	29	48	59	37	32	20	36	44
December	32	65	31	42	16	27	20	56	78	38	42	20	44	29
Average	46	66	36	42	21	37	33	41	62	32	29	18	28	111

to be closely connected with solar activity a similar connection of solar activity and radio signal strength was to be expected, and Pickard, in a recent paper,[5] has shown that such a relationship exists.

Rough observations on signal strength by the shunted telephone method were begun in this laboratory in 1915, and on the strength of atmospherics a little later, while since 1922 measure-

TABLE III

Average Signal Intensity and Atmospheric Disturbances for El Cayey (NAU) in microvolts per meter

	A. M.		P. M.	
1926	NAU	Dist.	NAU	Dist.
January	78	13	69	15
February	81	13	70	20
March	96	17	65	24
April	66	29	51	42
May	83	17	54	71
June	80	18	51	80
July	65	17	41	112
August	72	23	61	147
September	91	16	57	80
October	66	15	65	30
November	83	14	54	19
December	92	11	83	13
Average	79.6	16.9	60.1	54.4

ments of these quantities have been made which have considerable accuracy. These observations, which were originally planned for the purpose of making comparisons with other natural phenomena furnish considerable material for the present study.

In Fig. 5 the annual averages of sunspot numbers and the all daylight reception from Nauen in Washington from 1915 to 1926 is seen. The earlier years of the Nauen reception curve have little

[5] Proc. I. R. E., vol. 15, p. 83; 1927.

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claim to accuracy but it is certain that there was a reception maximum in 1917 and low values from 1920 to 1924. The signals are reduced to uniform antenna current.

Fig. 6 shows the relationship between the monthly average sunspot numbers and the monthly average strength of the all daylight signals from Nauen covering a period of five years from 1922 to 1926 inclusive. In order to eliminate the effects of seasonal variations on the signal strength, the observational points on the curve are the deviations from the means of the same months for the five years.

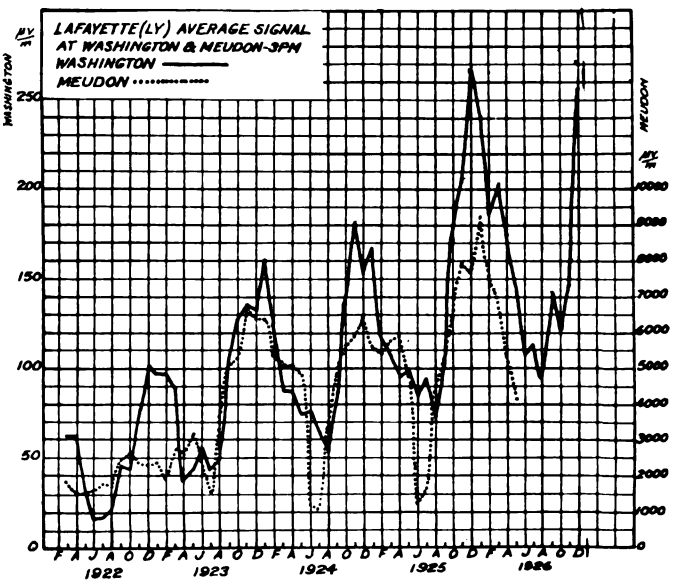

Fig. 4

Fig. 7 shows a similar curve for the averages of a number of stations, from 1924 to 1926. These curves of sunspots and signals, while not following each other exactly from month to month, show a quite definite positive correlation between solar activity and strength of long wave daylight radio transmission averaged over long periods.

Thus far our conclusions are reasonably certain, but in what follows I feel that we are on much less secure ground.

If the existence of a correlation between solar activity and radio be accepted, it is natural to suppose that there may be variations in radio reception which may show a connection with

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TABLE IV

MISCELLANEOUS OBSERVATIONS ON SIGNAL INTENSITY IN MICROVOLTS PER METER

1926

	AXL	AGX	GB	IDG	LCM	MUU	NPL	NBA	PCG	SAQ	YN	AFL	NPG	GKB	FZ	IDO
January	44.4	30.0	—	33.0	23.5	43.0	—	27.3	70.5	30.0	—	6.0	—	8.3	—	—
February	40.6	24.2	78.0	51.0	29.1	37.2	41.3	30.6	73.2	34.8	32.0	5.5	—	3.0	4.0	—
March	38.7	26.2	76.8	—	22.4	31.6	65.5	31.8	62.1	33.8	27.3	6.0	32.0	4.2	3.0	—
April	49.7	31.5	57.0	—	20.6	—	47.0	29.7	61.5	46.2	30.0	4.0	—	5.0	4.0	—
May	46.8	27.8	49.2	36.0	25.6	28.2	44.3	31.3	62.5	42.8	24.1	4.0	—	—	—	—
June	41.3	26.2	51.5	38.4	18.2	27.2	50.1	32.1	58.0	39.3	23.1	3.3	28.6	—	—	15.0
July	34.9	22.0	43.0	36.9	14.2	24.1	57.3	22.8	56.1	32.3	20.5	—	22.6	—	—	10.6
August	31.5	21.5	40.5	30.0	15.0	22.7	34.3	22.0	56.0	28.5	18.2	—	—	—	—	—
September	30.0	30.4	47.5	45.6	14.6	18.5	54.0	—	54.0	28.5	15.0	—	—	—	—	—
October	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
November	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
December	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
Average	39.7	26.6	55.4	38.7	20.3	26.5	49.3	28.4	61.5	35.3	23.7	4.8	—	—	—	—

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the period of rotation of the sun. It would not of course be expected that the changes in strength of reception would regularly follow the solar rotation, but if the irregular terrestrial factors could be eliminated, it might well be that, on an average, a waxing and waning of the signal strength would be found as the sun rotates.

When it is suspected that observed values may have a tendency to vary in a definite period, one of the best methods of handling the data is to divide the observations into groups of days equal to the suspected period, and then average the corresponding observations of the successive groups. By this process the real

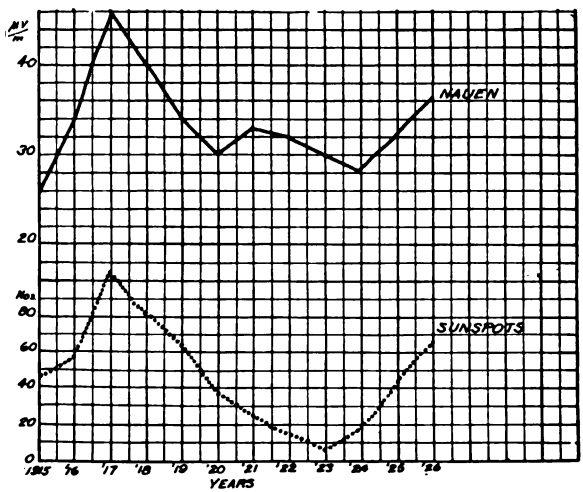

Fig. 5—Annual Average Signal Intensity of Nauen (AGS), 10:00 A.M. and Sunspot Numbers.

periodic changes should build up while others should tend to cancel. The apparent time of solar rotation is somewhat indefinite since the surface of the sun is gaseous. At the solar equator, the rotation period is about 25 days, at a latitude of 75 degrees, it is 32 days, while in the region where the sunspots are most numerous, it may be taken approximately as 27 days.

In Figs. 8 and 9 the signal field strength deviations from monthly averages have been divided into successive periods of 27 days, and the periods averaged, all the first days of all the periods, then all the second days and so on. Fig. 8 shows the results of this procedure for the sunspot numbers and for the daylight signals from Nauen AGS for the three years, 1922–1924 inclusive,

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and also for 1925 and 1926.[6] From this figure it appears that the Nauen signals, like the sunspot numbers, vary with the solar rotation and keep persistently in nearly opposite phase to them.

Some stations like LPZ (Argentina) appear to show very little 27-day variation; while in some cases, Bordeaux for example, Fig. 9, there seems to be a strange persistence in the signal curve forms even after an interval of a large number of 27-day periods. It is interesting to note in Fig. 9 that the two curves of the first halves of the two different years are more nearly alike than the first and second halves of the same year. In this connection it

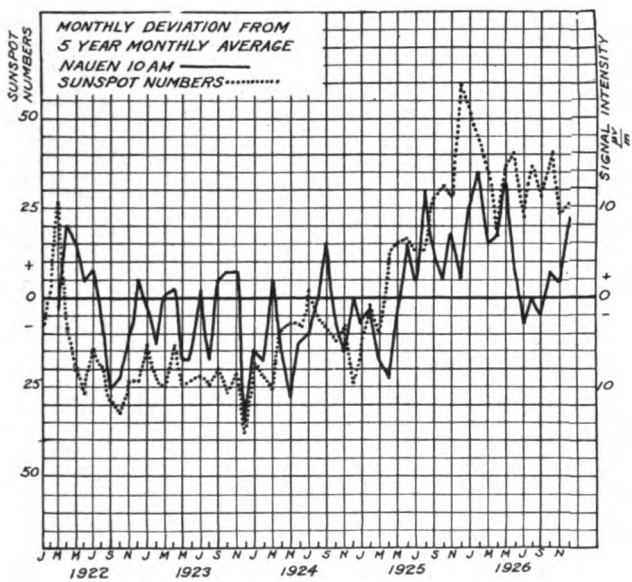

Fig. 6

may be remembered that the spots of the sun's northern hemisphere face the earth from December to June, while from June to December the southern spots face us. In addition to the 27-day periods there also seem to be indications of periodic changes in signal strength in shorter periods than 27 days. Some of the stations appear to show a 9-day period, which is one third of the period of solar rotation, while in some cases even shorter periods can be detected.

[6] For the sake of clarity the observation points in all the periodic average curves have been repeated beyond the 27-day period and in some cases smooth curves have been drawn through the irregular curves of the actual averages.

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It seems scarcely possible that such detailed resemblances as are shown in Fig. 9 can represent anything real in solar and radio relations, but while they are very possibly without real significance,

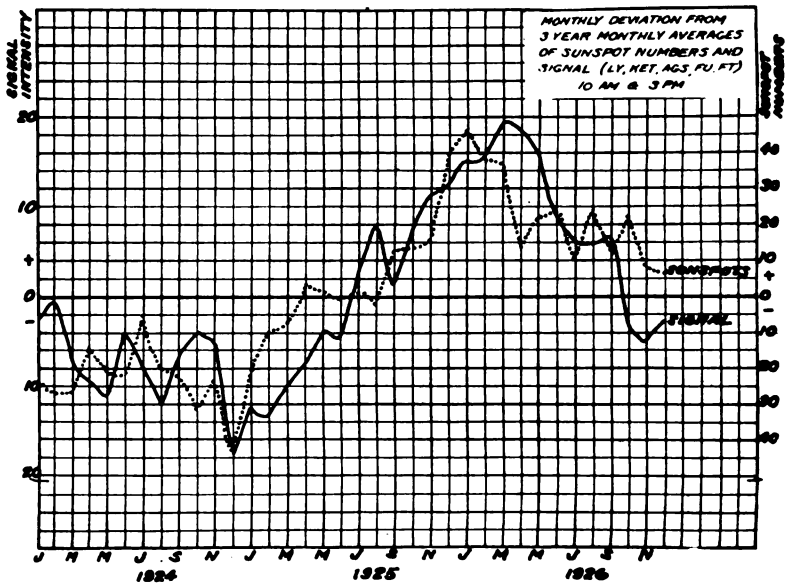
MONTHLY DEVIATION FROM 3 YEAR MONTHLY AVERAGES OF SUNSPOT NUMBERS AND SIGNAL (LY, KET, AGS, PU, FT) 10 AM & 3 PM
SIGNAL INTENSITY
SUNSPOT NUMBERS
SUNSPOTS
SIGNAL
J M M J S N J M M J S N J M M J S N
1924
1925
1926

Fig. 7

the similarities are tantalizing. The continuation of these periodic changes through several years, if they are real, would indicate fixed active areas on the sun which persist over very long periods.

TABLE V
DATA FOR TABLE IV.

Station	Approximate wave length	Frequency	Location
AXL	18 300 m	16.3 kc	Warsaw, Poland.
AGX	14 500 m	20.6 kc	Eilvese, Germany.
GB	7 800 m	38.4 kc	Glace Bay, N. S.
IDG	21 000 m	14.2 kc	Pisa, Italy.
LCM	12 000 m	24.9 kc	Stavanger, Norway.
MUU	14 000 m	21.4 kc	Carnarvon, Wales.
NPL	10 000 m	29.9 kc	San Diego, Calif.
NBA	6 500 m	46.1 kc	Darien, Panama.
PCG	17 800 m	16 8 kc	Kootwijk, Holland
SAQ	18 500 m	16.2 kc	Goteborg, Sweden.
YN	15 400 m	19.7 kc	Lyons, France.
AFL	7 800 m	38.4 kc	Königs-Wusterhausen, Germany.
NPG	10 500 m	28.5 kc	Mare Island, Calif.
GKB	6 800 m	44.0 kc	Northolt, England.
FZ	10 200 m	29.3 kc	Beirut, Syria.
IDO	11 000 m	27.2 kc	Rome, Italy.

In addition to the comparisons of solar activity and signal strength, a study is also being made of possible relations between solar activity and the strength of daylight atmospheric dis-

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TABLE VI
TRANSMISSION DATA FOR TABLES I, II AND III.
1926.

	Frequency	Wave Length	Antenna Current	Effective Height	Distance
	f	λ	I	h	d
					km
LY[a], Bordeaux	15.9	18 900	570	180	6160
FU, Ste. Assise, Paris	15.0	20 000	475	180	6200
FT, Ste Assise, Paris	20.8	14 400	380	180	6200
AGW[a], Nauen, Berlin	16.5	18 100	457	170	6650
AGS[a], Nauen, Berlin	23.4	12 800	398	130	6650
GBR[a], Rugby	16.1	18 600	653	185	5930
GBL, Leafield, Oxford	24.4	12 300	210	75	5900
LPZ, Monte Grande, Buenos Aires	23.6	12 700	600	143	8300
LPV, Monte Grande, Buenos Aires	17.0	17 600	565	143	8300
SPR, Rio de Janeiro	13.8	22 000	780	150	7800
KET[a], Bolinas, San Francisco	22.9	13 100	659	51	3920
ICC, Coltano, Pisa	19.9	15 000	380	150	7100
NAU, Cayey, Porto Rico	33.8	8 870	150	120	2490

[a] Daily antenna current reported. Other antenna currents more or less uncertain.

turbances. This will be more fully reported at another time, but the preliminary results are as follows: Here the connections are by no means so clear as in the case of the signals. A comparison of the

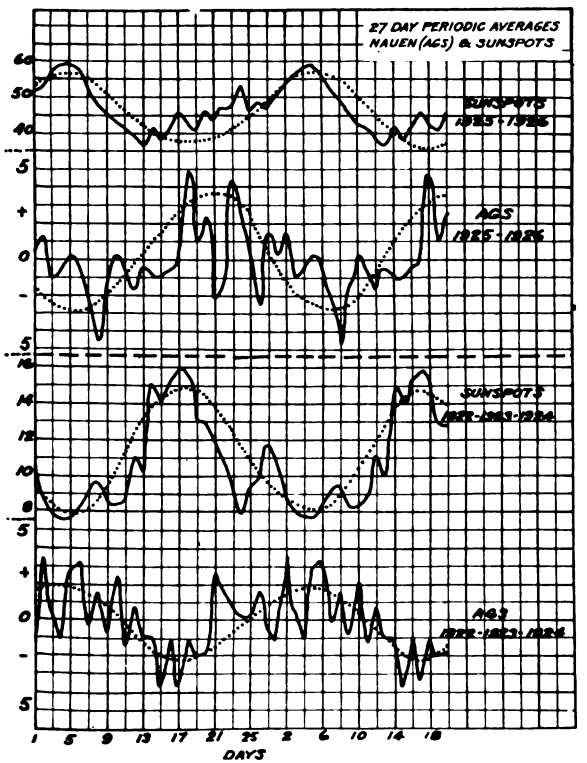

Fig. 8

monthly averages of sunspots and long wave daylight atmospherics, covering several years, shows little certain correlation

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between the two. If a correlation exists, it is apparently negative; that is, there has been a tendency since 1924 for the atmospherics to decrease with the increasing sunspot numbers of the advancing 11-year cycle. Twenty-seven day periodic averages also appear to show some degree of negative relation for the daylight atmospherics.

In conclusion, the observations show with considerable certainty that there is a general increase of signal strength with increasing sunspot numbers. There also appears to be a possible periodic relationship between the sunspot numbers and daylight

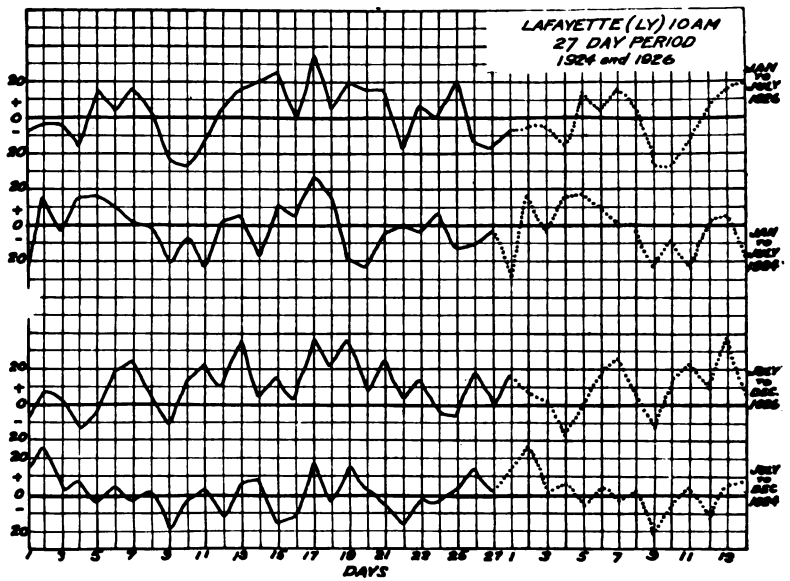

Fig. 9

signals, in which, in the case of most stations observed, the signals are nearly in opposite phase to the periodic changes of the sunspots. This is in agreement with the results of Pickard in the broadcasting range. While the work thus far must be considered to be in the preliminary stage, it seems probable that the relations of solar activity and radio phenomena will be found to be as worthy of study as those of solar activity and terrestrial magnetism.

A limited number of mimeographed copies of the Bureau of Standards record of daily signal measurements of long wave stations since January 1, 1924, are available for distribution to those engaged in the study of radio transmission phenomena.

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RADIO ATMOSPHERIC DISTURBANCES AND SOLAR ACTIVITY*

By

L. W. Austin

(Laboratory for Special Radio Transmission Research, Bureau of Standards)

The suggestion that atmospheric disturbances might be due to a bombardment of the earth's atmosphere by electrified particles from the sun was first made by De Groot[1] in 1917. A number of years later the U. S. Navy began experiments looking toward the establishment of a connection between these disturbances and solar activity. It was thought that there might be such a connection in the case of the type of atmospherics which sometimes produces simultaneous disturbances in the receiving apparatus

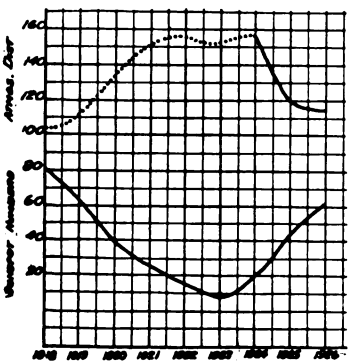

Fig. 1—Annual Average Atmospheric Disturbances, 3:00 P.M. (24 kc, 12500 m) and Sunspot Numbers.

at widely separated points as in Honolulu and San Francisco or even in Honolulu or San Francisco and Berlin.[2] There seemed to be some evidence that these simultaneous disturbances took place when the large sunspots were in the center of the sun's disk facing the earth. The observations were made in San Francisco and in

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Read before the American Geophysical Union (Section of Terrestrial Magnetism and Electricity), April 28, 1927.

* Published by permission of the Director of the National Bureau of Standards of the U. S. Department of Commerce.

[1] C. De Groot, Proc. I. R. E., vol. 5, p. 75; 1917.

[2] Baumler, Jahrb. d. Drahtlosen Teleg., vol. 19, p. 325; 1922. Proc. I. R. E., vol. 14, p. 765; 1926.

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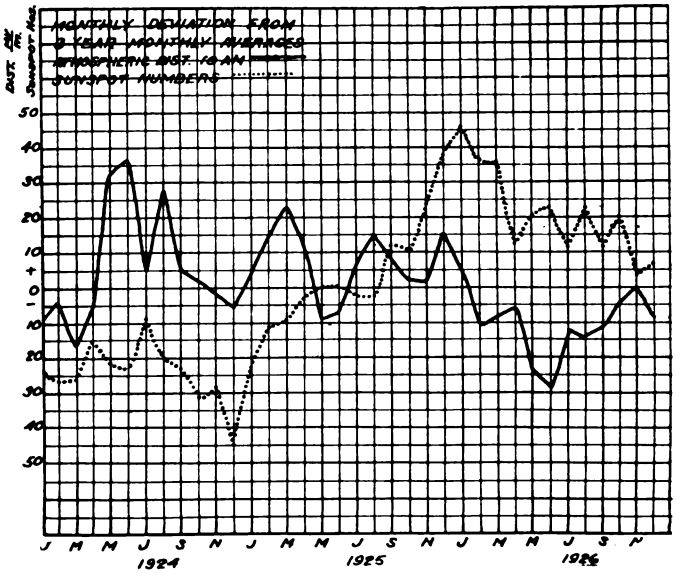

Fig. 2

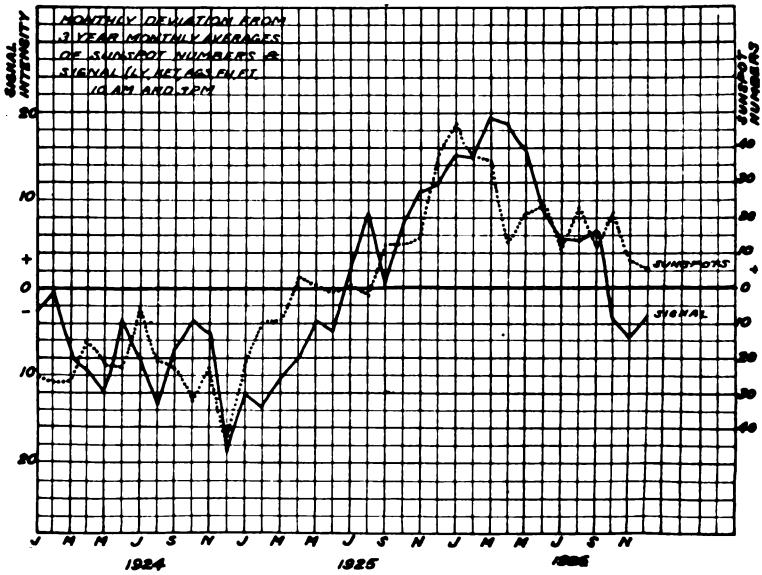

Fig. 3

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Washington and were continued intermittently for more than a year but without leading to any definite conclusion.

During the past year an examination has been made of the observational material on daylight disturbances between 9000 and 20,000 meters wavelength collected by this laboratory since 1918. In the earlier measurements the strength of the atmospherics was determined by the shunt across the telephone at which three crashes could be heard in ten seconds. This was subject to all the general inaccuracies of the shunted telephone method, including the effect of changing observers and possible changes in the sensitiveness of the telephones from time to time. The present

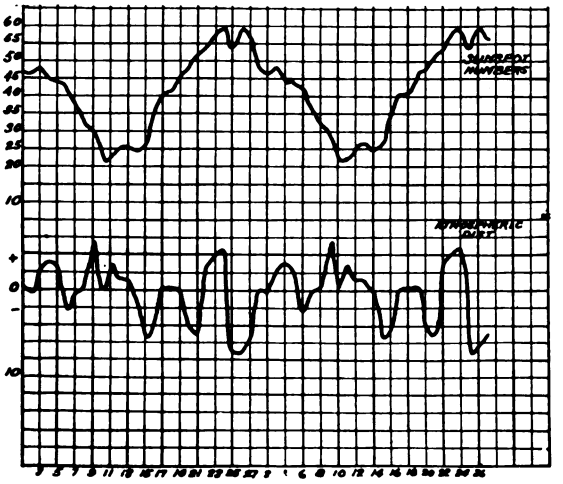

Fig. 4—27-Day Period. Atmospheric Disturbances 10:00 A.M. (f—15kc, 20,000 m.) and Sunspot Numbers, 1925.

method which was adopted in 1924 rates the strength of the disturbances in terms of the measured strength of an artificial signal which is just readable through them. The changes in methods of measurement do not permit comparisons of the earlier and later observations which are of much quantitative value. It seems certain, however, that the daylight atmospherics at 12,000 meters between 1920 and 1924 were considerably stronger than during 1918 and 1919 and during 1925 and 1926. This would indicate a general negative correlation with sunspots. This is shown in Fig. 1. The values before 1924 are dotted to indicate their comparatively low accuracy. Fig. 2 shows the deviations of the individual monthly averages from the three-year monthly averages

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of the 20,000-m. atmospheric disturbances and of the sunspot numbers from 1924 to 1926. The deviations are used, rather than the monthly averages themselves, in order to eliminate the large seasonal variations of the atmospherics. The increase in the sunspot numbers with the advancing eleven-year cycle is evident in the figure. The atmospheric disturbance curve, however, is rather noncommittal, but with some evidence of a negative correlation. The indistinctness of this relationship can be compared with the much more definite correlation of sunspots and signal intensity, as shown in Fig. 3, where the mean values of the daylight

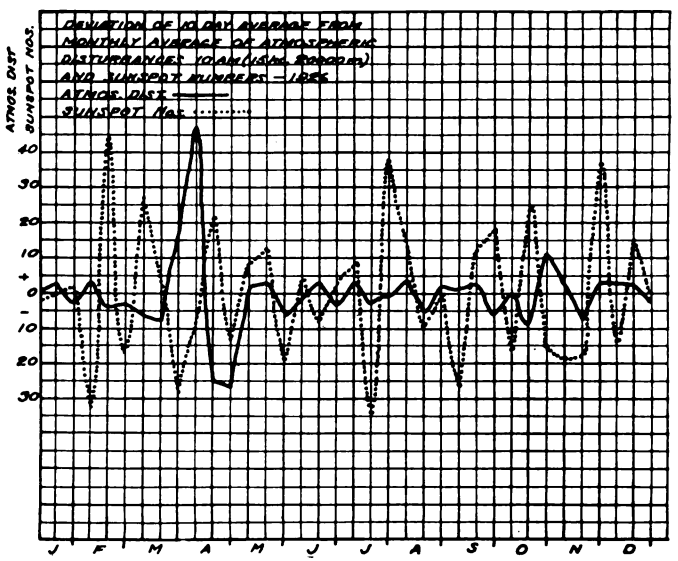

Fig. 5

signal strength of several long-wave stations are compared with the sunspot numbers by the same methods and over the same period of time.

If the atmospheric disturbances are dependent upon sunspots it would seem probable that there would be regular changes in disturbance intensity in the period of the sun's rotation. Fig. 4 shows the 27-day periodic averages of the 20,000-m. atmospheric disturbances and sunspot numbers in 1925, the points being repeated during a second 27-day period for clarity. The disturbance curve is somewhat smoothed by 3-day moving averages. It is seen that this curve, while very irregular, shows perhaps

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some slight evidence of being in nearly opposite phase to the sunspots.

Fig. 5 shows a comparison of 10-day averages of 20,000-m atmospheric disturbances and sunspots from January 1 to December 31, 1926. Here the prevailing correlation appears to be also on the whole negative.

For a comparison with these rather indefinite evidences of the solar influence on atmospherics, Figs. 6 and 7 show their much more definite connection with terrestrial phenomena. Fig. 6 represents the quite close relationship between 20,000-m. atmospherics

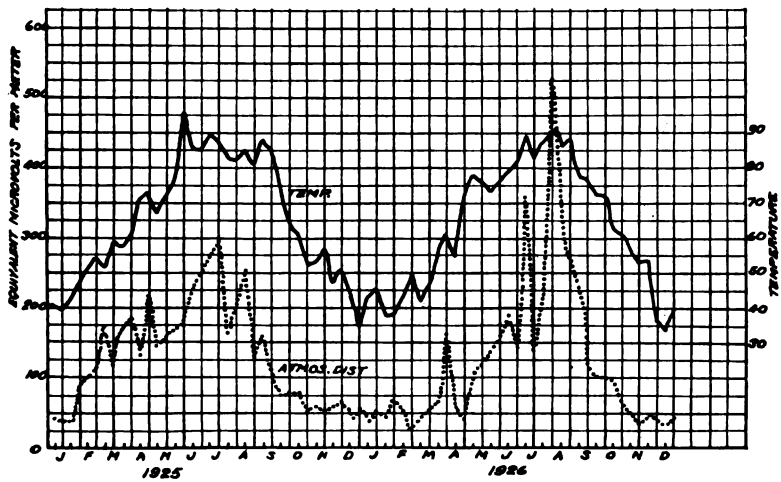

Fig. 6—Temperature and Atmospheric Disturbances—3:00 P.M. Ten Day Averages.

and local temperature averaged in 10-day periods. Here the well-known parallel variations with seasonal temperature are seen, and in addition there is a remarkable correspondence even in the small peaks and troughs of the two curves. Fig. 7 represents the relationship of atmospherics and the number of thunderstorms recorded in a region within approximately two hundred miles of Washington. Here again, as in the case of the temperature, we find a close relationship. The connections shown in Figs. 6 and 7 evidently have to do with atmospherics of comparatively local origin. Many of the disturbances are, of course, known to come from great distances, but these two figures show how prominent a part relatively local atmospherics play in the difficulties of radio reception in Washington.

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In conclusion, while there seems to be some evidence of solar influences on long-wave daylight atmospheric disturbances, at present, the proof seems insufficient to establish this with certainty.

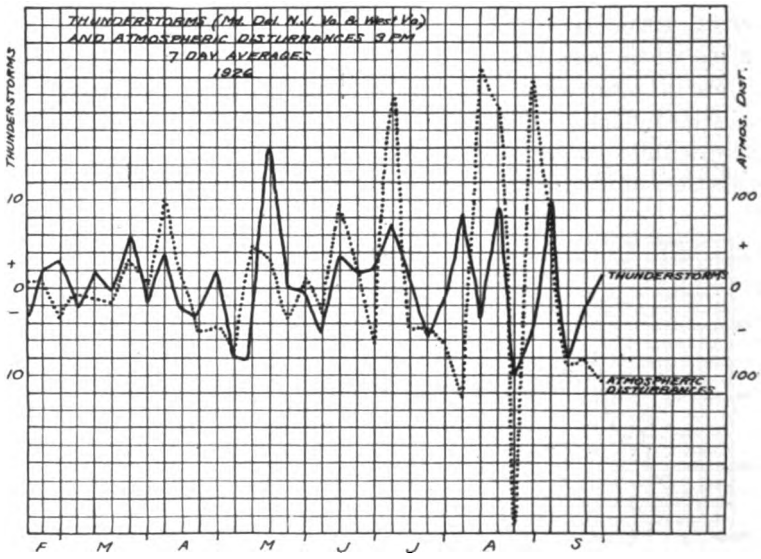

Fig. 7

It may be that the influence of solar activity on the weather and that of the weather on the atmospherics may be the indirect path by which the connection must be traced.

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TWO CONTRASTING EXAMPLES WHEREIN RADIO RECEPTION WAS AFFECTED BY A METEOROLOGICAL CONDITION*

By

LIEUTENANT E. H. KINCAID, U.S.N.

(Hydrographic Office, Navy Department)

The writer has had opportunity at sea to study static with relation to the atmosphere since the spring of 1924 on board the U.S.S. *Kittery*. During 1926, these observations were made with the assistance of the radio compass. The research originally started on board the U.S.S. *Kittery* is now being continued by the Hydrographic Office, Navy Department, in collaboration with the Bureau of Engineering, Naval Communications, and the Bureau of Aeronautics.

The purpose of this article is primarily to show that static has sufficiently definite relationship to the distribution of the atmosphere as plotted on the daily weather map to enable one by proper observations to make use of static in weather forecasting, and to make use of our present knowledge of atmospheric distribution and movement in static forecasting.

This problem must be considered from two different viewpoints.

The first is by making use of instantaneous static intensities without reference to any directional properties that static may have. This might be called plotting the Highs and Lows of static such as we now plot Highs and Lows of barometric pressure.

The first experiment conducted by the U.S.S. *Kittery* for the Navy Department before the introduction of the direction finder was on this principle.

Considering that a definite relationship is established between atmospheric structure or changing structure and static, facts ascertainable concerning static from one or more points will be of value in forecasting weather and our knowledge concerning development and movement of atmospheric conditions will assist in making static forecasts.

Received by the Editor, June 8, 1927.

* Presented before the American Section, International Union of Scientific Radio Telegraphy, April 21, 1927.

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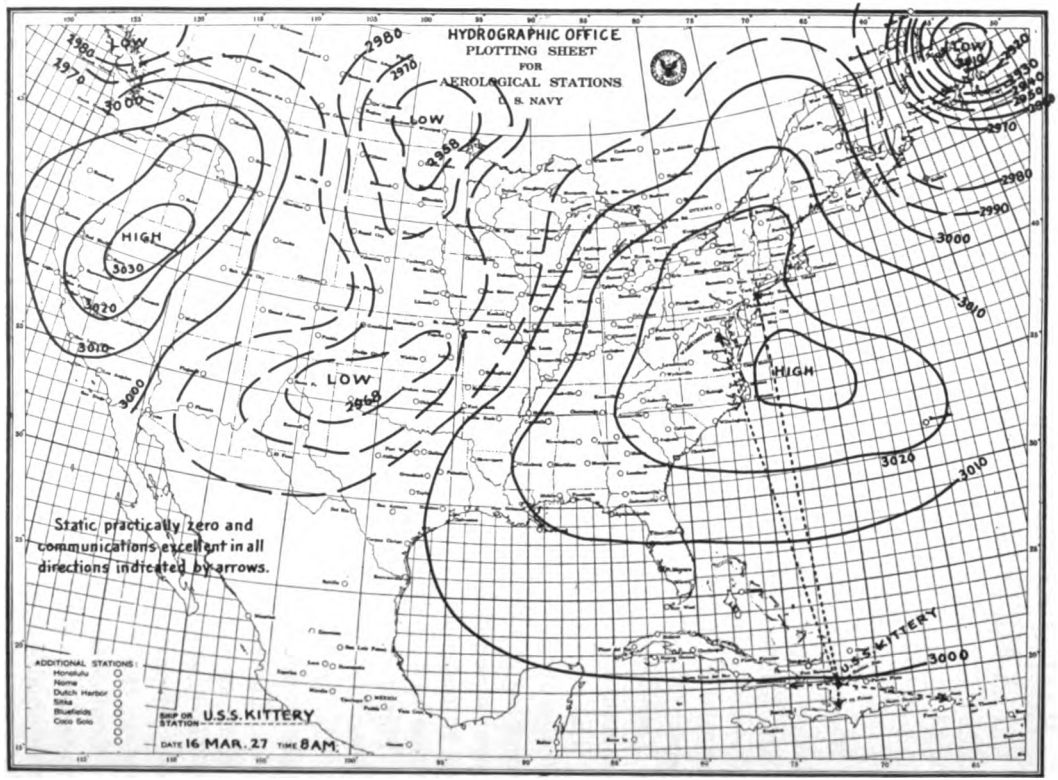

Fig. 1

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Add to this knowledge what directional properties static is found to have and we begin to realize the possibility of the value of a static observation to a condition remote from the station making the observation. It is this latter feature that will prove of greatest advantage to the mariner.

Two examples are given herewith of well-known meteorological conditions with contrasting types of pressure distribution. One is a well-formed homogeneous High which covers the West Indies, the South Atlantic and East Gulf States, and the ocean area adjacent thereto.

Fig. 1 illustrates the static-free and excellent communication conditions attending this High. The ship at the time off Cape Haytien, Haiti, on 16 March 1927, with a two-kilowatt set had no difficulty in communicating with Port au Prince, Haiti, Guantanamo, Cuba, San Juan, Porto Rico, Arlington, and New York.

Fig. 2, the other example, is of the same geographical area but with a hurricane, the most intense type of Low (save tornadoes) occupying a part of the same area. The Low in this case is the historical Miami hurricane shortly after it had passed over and devastated Turks Island. The position of the observer was the same during the observance of both states of weather and attendant radio reception conditions, that is, off Cape Haytien, Haiti, in this case on the evening of 16 September 1926.

In the former, the High, static was practically zero and reception excellent in all directions over long ranges within the High.

In the latter, the Low or Miami Hurricane, reception was impossible although attempts were made to communicate with all stations indicated by arrows in Fig. 2, namely, Arlington, San Juan, Guantanamo, Port au Prince, and even Cape Haytien, Haiti, which was only 40 miles distant.

Our steering gear broke here as we were entering Tortuga Channel to obtain a better lee, and had an SOS been necessary, it would have been drowned out by the intensity of the static. Fortunately, the U.S.S. *Kittery* is a twin screw ship, and we used our engines until the steering gear was repaired.

Here we deal not with theory but with facts, namely, (1) that on one occasion static was terrific and the observer was within the 29.60 isobar, southeast quadrant of a hurricane, and (2) on the other occasion the observer was within a homogeneous High and static was practically nil.

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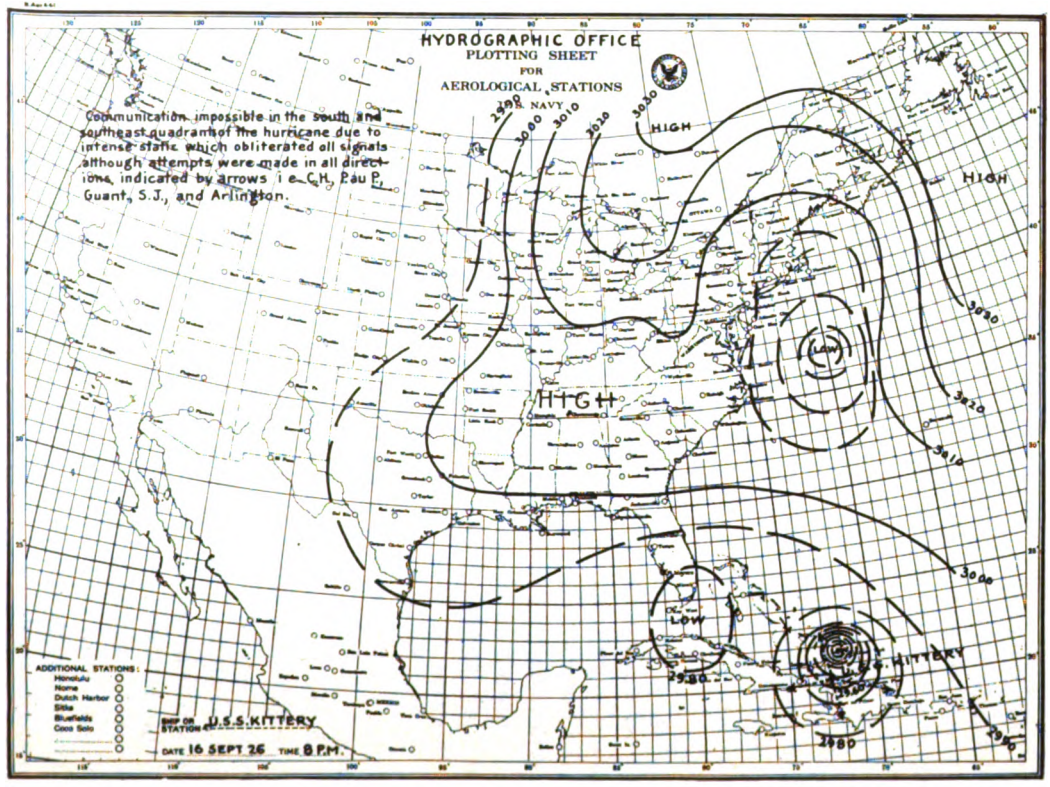

Fig. 2

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The writer wishes to ask what this and similar phenomena would seem to indicate. To the observer in this case the same relative situations have recurred so often that he is of the opinion that Highs are relatively static free and Lows are attended with static somewhat proportional to the intensity of the disturbance. This seems to hold good from troughs of Low pressure to hurricanes and from poorly defined Highs with mild static to great homogeneous Highs with practically no static.

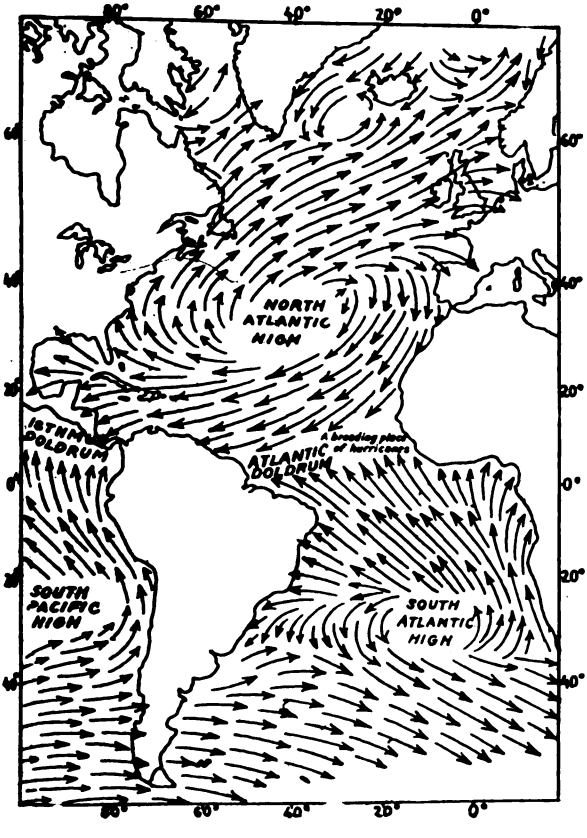

Fig. 3—Breeding Place of Hurricanes—Doldrums.
(After Bartholomew's Physical Atlas, Vol. III, Plate 14.)

This High had traveled across the United States from the Pacific States where it was central on 11 March. We have every reason to believe that the same relative good reception condition had identical motion of translation.

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The rate of the motion of translation of Highs is shown in many publications of the U. S. Weather Bureau and authors of meteorological works. (See Fig. 14.)

With the same accuracy as Highs are now used in weather forecasting they can be used in forecasting the negative information concerning static conditions.

The Low, or Miami Hurricane, had traveled at the rate of about 300 miles per day from the doldrums well out to the eastward of the Lesser Antilles, possibly near the Cape Verde Islands. It is a certainty that this area of heavy static and impossible communication conditions had traveled at the same relative position to the hurricane and at the same rate of 300 miles per day. (See Fig. 4.)

Considering that these things are true and that Highs and Lows and our knowledge concerning their development, disintegration, and motion of translation is to be the basis for our static forecasting, we need next to study the structure of Highs and Lows with a view to locating that part of each which may under the most general conditions be expected to give the greater amount of static. We also want to know where within the High and Low to expect the minimum amount of static for that particular structure.

In the Miami hurricane, static was stronger in the southeast quadrant than in the west and southwest quadrants. In general, the eastern half of Lows will be more heavy in static than the western; the southeast, in general, than the northeast; and the southwest, than the northwest.

On 21 March when the High mentioned above had moved to the eastward until the *Kittery's* position was in the southwest quadrant, static increased steadily.

In Highs it has been my observation that the center has the mildest static, the southeast quadrant less than the southwest, the northeast less than the northwest. Especially is this true near the edge of the structure.

Highs of different origin and hence different structure will have different static conditions associated with them. A Pacific High, deep but of limited area, will have a different static association from an Alberta High which is shallow but of great magnitude. And similarly Lows of different origin will have correspondingly different static attendant thereto.

The above facts can be reconciled with results obtained by many who are studying static with the view of correlating it with the atmosphere.

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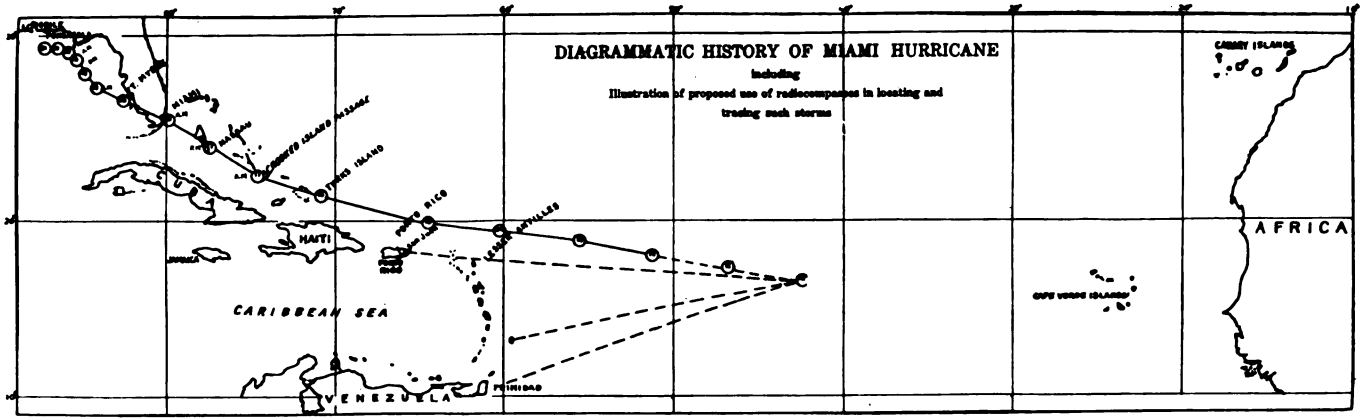

Fig. 4—Extending Known Path of Miami Hurricane Through Doldrums to Easternmost Breeding Place of Hurricanes. Known Part of Path Shown by Solid Line.

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A few observations of others who have studied this subject abroad may be added here:

Abstracts of Papers on the Meteorological Relations of Atmospherics

In June, 1923, the Council received a letter from Mr. R. L. Latcham, of the cable steamer *John W. Mackay*, Nova Scotia, calling attention to the possibility of using, for the purpose of weather forecasting, observations of the character of those naturally occurring electro-magnetic waves, called atmospherics, which are so familiar to all wireless listeners. The Council appointed a small committee to consider whether observations on atmospherics might be of service in forecasting. As a preliminary to the work of the committee, the writer undertook to prepare the bibliography of the meteorological relations of atmospherics which is now reproduced here.

The correlation between the occurrence of actual thunder or lightning and the radiation of atmospherics is so well-known that many references to it have been omitted, but it is believed that no other well-authenticated data have been overlooked.

It will be seen that very acute divergences of opinion exist as to possible relations between the intensity of atmospherics on the one hand and most of the phenomena definitely recognized as "meteorological" on the other. There is an overwhelming consensus in favor of the obvious correlation with thunder, moderate agreement on the correlation with convective processes in the absence of reported thunder, but beyond this point, the evidence is mutually contradictory. Reasons for this are not difficult to assign. It is established beyond doubt that the range of reception of atmospherics may reach one or two thousand miles, it is most probable that the range frequently, if not usually, attains the length of the earth's semi-circumference. Comparisons between received atmospherics and local weather will therefore, in part, be a comparison of the weather of a parish with the electrical phenomena of a hemisphere. Moreover it is still impossible to find a scale and a classification for the intensity and character of atmospherics which shall be generally acceptable, unambiguous, and capable of assessment by the average observer. The Beaufort of atmospherics is not yet.

Nevertheless, the evidence that the atmospheric was well, if rashly, named is accumulating rapidly, and the summaries of the most recent work show that the location of "cold fronts" by radiotelegraphic observations on atmospherics is an established possibility. R.A.W.W.

After describing typical X effects from lightning, remarks that X's are more frequent in summer and autumn than in winter and spring, near high mountains than in open sea, in south than in north winds (in the Mediterranean), in the front of a cyclonic disturbance than in the rear, with falling barometer than with rising barometer.

Notes case of a heavy winter gale, without observed lightning, which was preceded by X disturbance.—Jackson, H. B.—London, Proc. R. Soc., 70 (A), 1902, p. 266.

X's in Mediterranean are worst when pressure is low, temperature high, and humidity low. The sirocco forms an exception to this rule, as it always brings atmospherics.—Crawley, C. G.—London, Elect., 69, Jan. 31, 1913.

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Describes observations showing that every observed lightning flash produces an X, multiple flashes producing multiple X's, and discusses the adequacy of the known thunderstorm distribution to account for all X disturbance. The author proposes direction-finding for the location of X sources.—ERSKINE-MURRAY, J.—London, Elect., 67, 1911, p. 219.

Describes the recording of thunderstorms up to 500 km. distance by crystal detector and galvanometer.—FLAJOLET, .—Paris, C.-R. Acad. Sci., 154, 1912, p. 729.

On the basis of one year's observations at Anche (Indre et Loire) the relations between X's and weather are thus enunciated (the original French descriptions being retained to avoid ambiguity by loss of onomatopoeia).

(1) Violent "craquements" indicate a thunderstorm, approaching if the X's become more and more frequent, receding if they become less frequent or less strong.

(2) A slight "sifflement" is produced by a heavy hailstorm passing near the receiver.

(3) Dry, isolated, and rather weak "claquements" generally precede a fall of temperature, a spring frost.

(4) If the wind is about to turn, the X's are of short wavelength, and seem to come in strings.

(5) Numerous "crepitements," accompanied by fairly regular strong and "fusant" "craquements," precede great barometric depressions and foretell gales.—FRANCK-DUROQUIER.—Nature, Paris, 41 (1), 1913, p. 218.

Finds that X's and changes of recorded potential gradient go together. Balloon observations show great increase of X's inside cloud, and a rapid decrease of X's with increasing height.—LUTZE, .—Phys. Zeit., Leipzig, 14, 1913, p. 1194.

Observations at Mt. St. Aubert, Tournai, May–September, 1912, led to the following conclusions: Absence of X's characterizes approach of fine weather; little precipitate cracklings—hail or heavy rain; distant cracklings—fine weather; numerous and prolonged cracklings—stormy weather; prolonged sound like water from a gutter—sudden change of weather; prolonged and singing sound seeming distant at first, approaching little by little till strong—violent thunderstorm or tempest.—DELVAL, —Quoted in Perret Maisonneuve's "La T.S.F. et la Loi," (1914).

An outburst of X's follows a sudden change of wind on the North Atlantic coast of the United States, especially a change from South to North. X's also accompany a rise of temperature there.—MARRIOTT, R. H.—Proc. Inst. Radio Engin., New York, 2, 1914, p. 37.

X storms coincide with convective weather, rapid fluctuations of pressure, and rapid movements of depressions.

In Malta, X's precede by several days the advent of convective weather.

In Australia, day-time rain is preceded by X's in 80 per cent of all cases.

In Ireland, X's are bad with a North-West wind on the Atlantic coasts.

In Sierra Leone, the periodic dry wind suppresses X's.—RADIO TELEGRAPHIC INVESTIGATIONS COMMITTEE OF THE BRITISH ASSOCIATION. London, Rep. Brit. Assn., 1915.

From observations in the Belgian Congo, 1915–17, the author believes that X's from lightning have an effective range not exceeding 100 or 200 km.

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In general a day of thunderstorm is preceded by several days of increased X disturbance.

In general a day of thunderstorm is followed by several days of reduced X disturbance. This law is more general than the former.—GOLDSCHMIDT, R. —"La Telegraphie sans Fils au Congo Belge," (Brussels, 1920).

Low X intensity is found under rainless cloud, X's increase as cloud begins to break up or as rain begins. Diminishing convection current goes with low X intensity.

From a comparison of annual variation curves the author summarizes as follows:—X maxima go with low potential gradient, but high convection current and high numbers of ions per c.c., and conversely for X minima.

From diurnal variation curves:—X's are bad when the vapor pressure is at a maximum at the surface and increasing above, when the conductivity at ground and aloft is increasing, and the surface wind decreasing. X disturbance is at a minimum with maximum conductivity at a moderate height, with overcast sky, with maximum relative humidity, minimum temperature, and maximum air pressure.—WIEDENHOFF, S.—Jahrb. drahtl. Telegraph., Berlin, 18, 1921, p. 242.

Increased X frequency indicates an imminent thunderstorm; if the intensity of X's is great, the thunderstorm will be violent, with wind and rain, if the X intensity remains low, the thunderstorm is feeble. If the intensity augments, the storm is approaching, and vice versa.

This diagnosis is valid at great distances, before the cloud associated with the storm has appeared at the receiving station.—MARCHAND, — Rev. Gen. Sci., Paris, 32, 1921, p. 594.

Infrequent but sustained X's with an intense high metallic note are due to more or less distant cloud. Visible lightning, and hail, give this type of X.

Lacosts (q.v.) is summarized.—ROTHE, E.—Paris, C.—R., Acad., Sci., 175, 1921, p. 840.

Enunciates laws based on radiogoniometric observations at Strasbourg.

(1) If a distinct well-developed depression exists, with closed concentric isobars, the maximum X disturbance proceeds from the south or south-east of the depression, the displacement in direction and variation of intensity allow of following the course of the depression.

(2) If the depression has a distant center and less curvature, the X maximum is still south or south-east of the periphery, but is less sharply marked.

(3) Secondary depressions, with "barometric pockets" and cols, correspond to storm fronts, the direction of X maximum is difficult to find, and X's are violent from all azimuths.

(4) A near thunderstorm gives violent X's from its azimuth.

Many X's received in clear weather are comparable in intensity to those of observed lightning. The author concludes that X's are due to more or less distant storm phenomena, and are often produced in overcast regions.

The last few days' work in the period in question indicates that fog facilitates X transmission.—LACOSTE, J.—Paris, C.—R. Acad. Sci., 175, 1921, p. 843.

X's come from definite azimuths when storm clouds are distant, but are observed on all azimuths when the storm is very near.

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Storm clouds produce X's which gradually cease as soon as uniform rain starts, but X's persist to some extent when rain is violent.

Strato-cumulus most often furnish well-defined and oriented rumbling X's.

Re-summarizes Lacoste, Q.V.—Rothe, E.—Ann. Phys., Paris, 17, 1921, p. 385.

Describes further work at Strasbourg in summer of 1922, confirming previous conclusions, and demonstrating possibility of forecasting new depressions. Remarks on the special value of the radiogoniometer in thunderstorm forecasting.

Some thunderstorms give X's which are numerous only on short wave observations.—Lacoste, J.—Paris, C. R. Acad. Sci., 176, 1922, p. 707.

Preliminary results of directional observations on atmospherics during 1916–18, by stations on British coasts, show that on an average five apparent sources of atmospherics per week were located. Of those checked against meteorological data, 15 per cent agreed with reported thunderstorms, 10 per cent with squall phenomena, 69 per cent with rainfall within the 24 hours containing the time of observation, without reported thunderstorms or squalls. Thus 94 per cent of the apparent sources were associated with rainfall. Thirty-five per cent of the cases examined fell on the forward edge of a rain area.—Watson Watt, R. A.—Nature, London, 1922, p. 680.

Gives results of observations at Heidelberg, September 1919–August 1921, on 2,000 and 12,500 metre wavelengths. In the warm half-year X's developed on the appearance of a new depression over North-West Europe, disappeared on its departure or filling up.—Wolf, F.—Jahrb. drahtl. Telegraph., Berlin 19, 1922, p. 289.

Observations in the plains of Upper Rhineland, 1915–17, on 600 to 2,000 meter wavelengths, led to a classification of X's in ten groups.

(1) Prevailing weather type, a whistling sound decreasing in strength, ending almost in a whisper.

(2) Must type, weak hissing sound.

(3) Cloud break-up type, repeated groups in the form of explosive crashes. Specially marked on break-up of alto-cumulus.

(4) Cirrus type, "trrrrtasss" (palatal r.), increasing as cirrus gathers, continuing till cirrus ceases to increase.

(5) Lightning type "rrrrrrssss" (palatal r.), a more rapid "r" succession than in (4).

(6) Front type (squally wind), creak like a revolving clapper.

(7) Cumulus-development type, short single sharp cracks.

(8) Thunderstorm-cumulus type, sharp cracks of different strength, coming in groups.

(9) Sunrise type, violent rattling and scratching of longer duration, not in group form.

(10) Sunset type, superposition of (3), (9) and other types, partly due to surface cooling and condensation, partly to cloud break-up.

Every stage in the development of a thunderstorm released a definite type of disturbance. Groups 4 and 9 are superposed in thunderstorms.

The lightning type was observed at night with perfectly clear sky.

The increase in disturbance numbers started several hours before the onset of the thunderstorm, though there were exceptions to this rule.

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In spite of unmistakable signs of thunderstorm the number of disturbances remained normal until a sudden release occurred.

Strong disturbances were found to come from strata of misty haze and also from forests with their well-known summer cloud formation.

Direction-finding differentiated between sharply directed X and those showing little directivity. Localized thunderstorms and squalls could be located in azimuth.—STOYE, K.—Jahrb. drahtl. Telegraph., Berlin, 20, 1922, p. 303.

Summarizes work on X's under the Meteorological Office and the Radio Research Board, gives examples of the location of thunderstorms up to 2,400 km. distant, by direction-finding on X's, and discusses the relations of X's, rainfall, and lightning.—WATSON WATT, R. A.—London, Wireless World and Radio Rev., 12, 1923, p. 601.

Atmospherics are the phenomenon that best gives evidence of the passing of meteorological disturbances in tropical regions; the other meteorological variables only indicate that passing in a much less regular way; moreover, they only give evidence of the meteorological disturbance when it has reached the observing station, while atmospherics announce it some hours in advance. —BUREAU, R., and others.—C.R. Acad. Sci., 178, 1924, p. 556; 1623; 179, 1924, p. 394; 180, 1925, p. 529; 1122; Onde, Elec., 4, 1925, p. 31; 58.

Observations at Lausanne 1919 to 1922, at Zurich 1924.

Preliminary conclusions are stated thus:

(1) Intensity frequency nature and direction of X's vary with altitude.

(2) This variation is closely connected with variations of vertical temperature gradient. Cooling augments, heating diminishes the phenomenon.

(3) Lower layers are much more disturbed than higher.

(4) Apart from X's of proved distant origin, range of X's originating in the lower layers does not seem to exceed 250 K., and remains almost always under 150 K. in mountainous regions.—LUGEON, J.—Paris, C.R. Acad. Sci., 180, 1925, p. 594.

Shows correlation between apparent sources of atmospherics and their meteorological environment. Of 490 cases examined, 25 per cent fell within 250 K. of reported thunder, a further 28 per cent were associated with more distant thunderstorms or with squall phenomena, 21 per cent more with rain, 13 per cent uncorrelated.

Reports the tracking of a cold front for 2,700 k. by automatic recorders of X's.—WATSON WATT, R. A.—London, Proc. Phys. Soc., 37, 1925.

Most of them seem to have correlated static with a local condition. I am trying to show how those local conditions are carried along with the general drift of the atmosphere and how static from them can be used in weather forecasting.

I have listed the above facts in this particular manner in order to show how much can be forecast concerning static once its relationship to High and Low structure is well established. From our knowledge of the development and motion of translation of Highs and Lows we have the basis of a static forecast which will

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be enhanced by an instrument which will make use of the directional properties of static.

Fig. 5 shows the Miami hurricane as it was passing over the Florida Peninsula.

Tracking the Miami Hurricane

Here it will be seen that from Guantanamo, where the U.S.S. *Kittery* was then moored, the heaviest static was in the general direction of the southeast quadrant. The center of the storm was 600 miles distant. (See Fig. 5.)

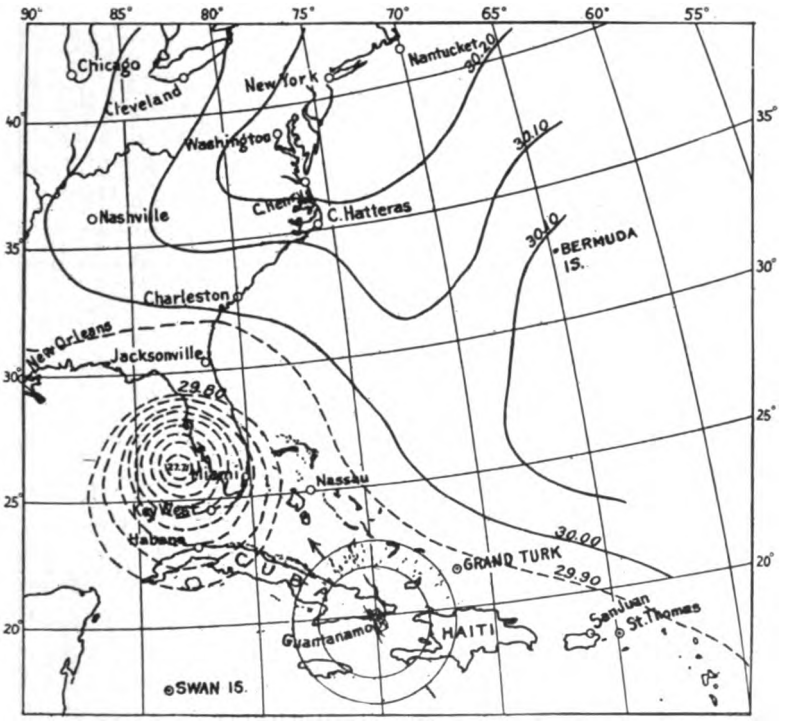

Fig. 5

Fig. 4 shows the location of the centers of the Miami hurricane from the time it was first observed to the eastward of the Lesser Antilles to the time it filled up and dissipated over Mobile, Alabama. Did not this static belt drift along with the hurricane? How far will a direction finder be able to reach out with sufficient precision as to make its use practical in habitually tracking such storms? The dotted lines show where the system may enable us to reach out in the region to the eastward of the

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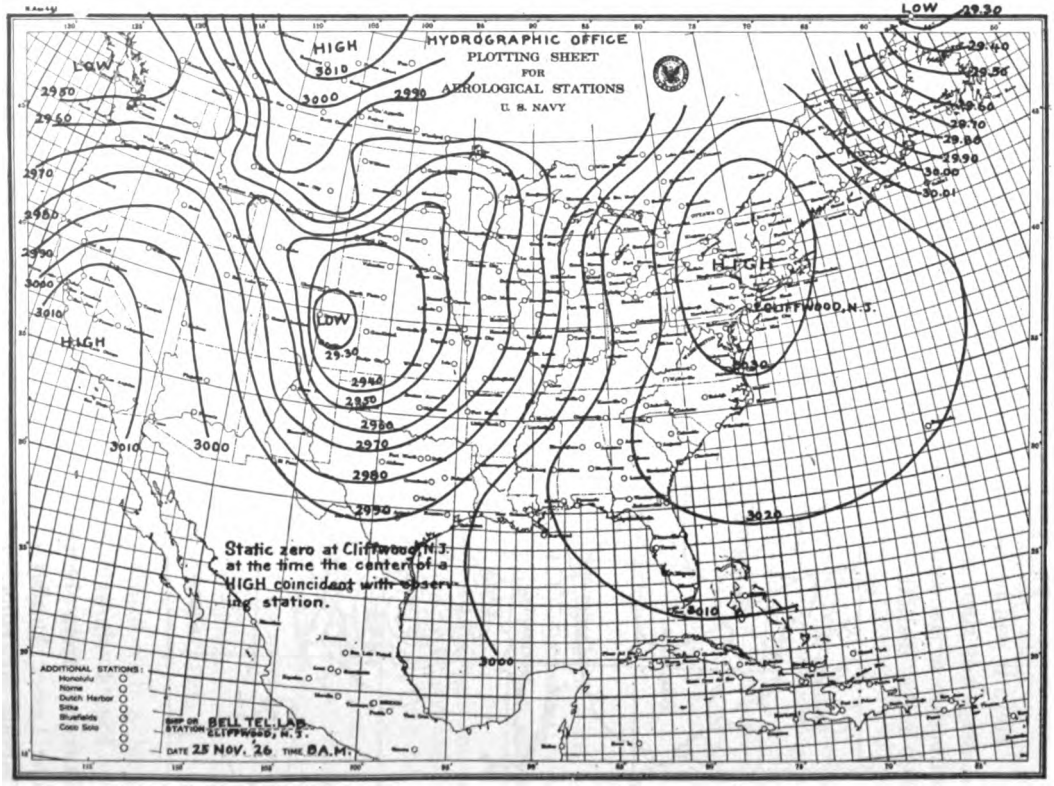

Fig. 6

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Lesser Antilles where barometer readings are scarce and triangulate such a static center in hurricane season, from St. Thomas, Barbados, and Trinidad.

The Doldrums' Breeding Places of Hurricanes

Fig. 3 shows the ideal atmospheric distribution for summer, from which the hurricane belt and the doldrums may be studied. In addition to those hurricanes that form to the eastward of the Lesser Antilles, we see the breeding place for the early spring and late fall type that form around the Isthmus of Panama, between the North Atlantic or States High and the South Pacific High. The selection of compass stations here as elsewhere will depend upon the proved range of accuracy of the radio compass for this kind of work.

Anyone working a field as new and undeveloped as this naturally finds comfort in another independent source wherein his conclusions have been verified. Mr. Friis of the Bell Telephone Research Laboratory has taken observations with an instrument developed at that laboratory over a period extending from last August up to the present date. They very generously coöperated with the Hydrographic Office in this work and I show you here the results of a part of their observations which substantiate work previously done by me.

Fig. 6 shows the High central at Cliffwood, New Jersey, November 20, 1926. Static was zero during the passage of this homogeneous High.

Following its passage, the Low shown in Fig. 7 in its motion of translation to the eastward passed over the same station. Static rose in intensity from 1 to over 400 microvolts per meter during its passage.

The direction shifted consistently from west as shown to north of east during the passage.

The graph of direction and intensity of static shown in Fig. 9 was received by their machine.

Some idea of the effect of this static intensity on the received signal is gained when the reader realizes that the Bell Telephone System try to maintain for trans-Atlantic work a received signal strength for code of five microvolts per meter and for telephony of 15 to 20 microvolts per meter.

The Hydrographic Office is pleased to see their desire of a year ago so quickly fulfilled by commercial enterprise. We said then that an instrument that would record both direction and intensity

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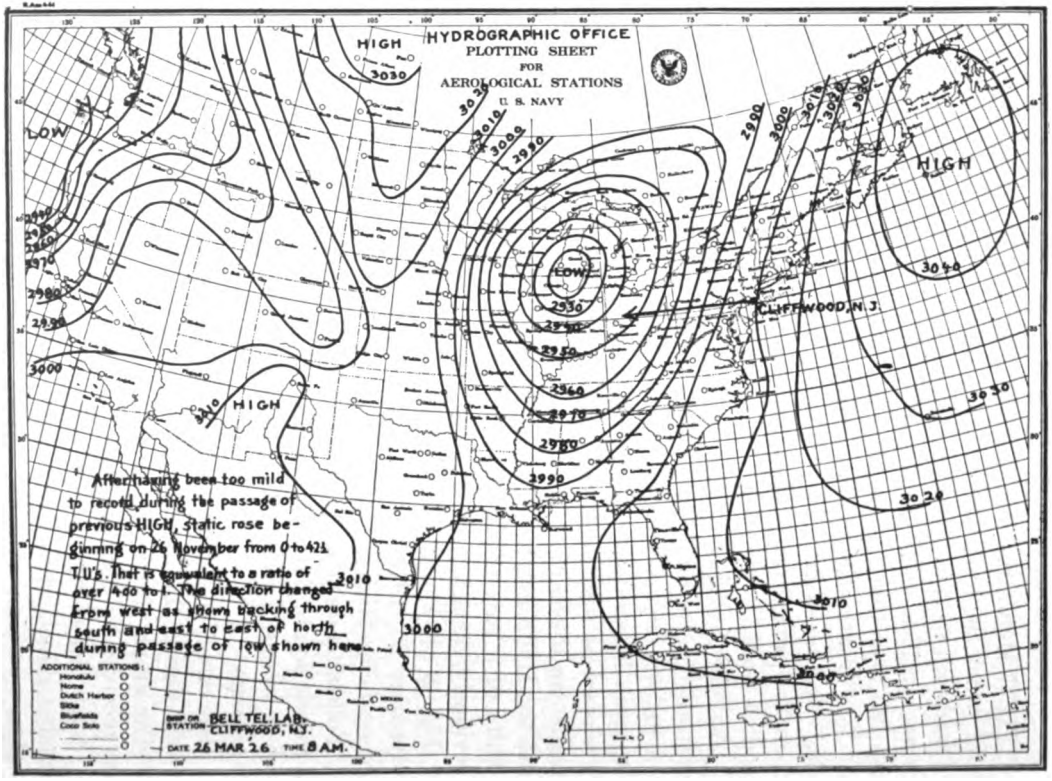

Fig. 7

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of static would be necessary to the successful accomplishment of this work. Dr. Austin at the Bureau of Standards has also developed a fine automatic static recorder and there are still others.

The recorder we used is shown in Figs. 8 and 15. It is slightly different from the hookup we used at Bellevue Laboratories which was as follows:

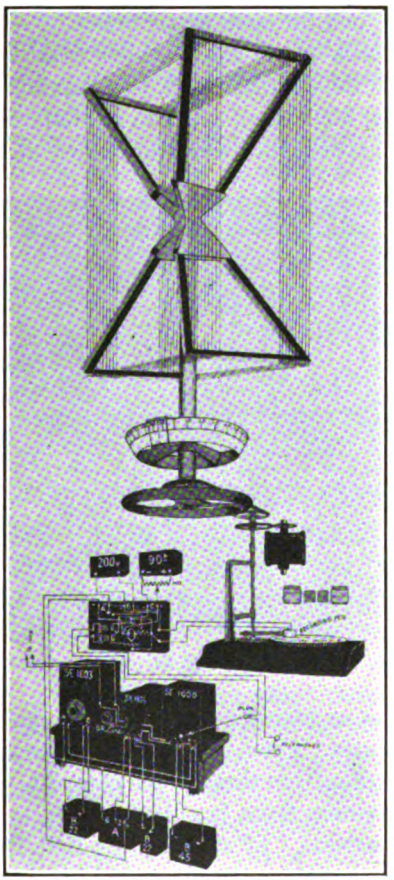

Fig. 8—Kincaid Static Recorder

The loop, an (SE-515A), was 6 ft. high, 4 ft. 3 in. wide, and consisted of 90 turns, spaced over an axial length of 4 ft. 3 in. The loop center was grounded.

The terminals were connected to an (SE-1834) amplifier and tuned by a 0.01 μf. variable condenser (SE-1830).

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The amplifier had three radio frequency stages, a detector, and two audio frequency stages of amplification.

An additional audio frequency stage was used, the output of the latter being connected in parallel, and the negative grid bias being adjusted for zero plate current, at zero signal, by means of a potentiometer. The plate current of the two tubes actuated a pen which was shunted by a 1 μ.f. condenser.

The instrument combination described above results in depression of the pen for each static crash.

The pen records on a card, which rotates very slowly in synchronism with the compass loop, two maxima and two minima being indicated for 360 deg. of rotation (a bilateral loop operation). Some unilateral experiments were conducted at Bellevue.

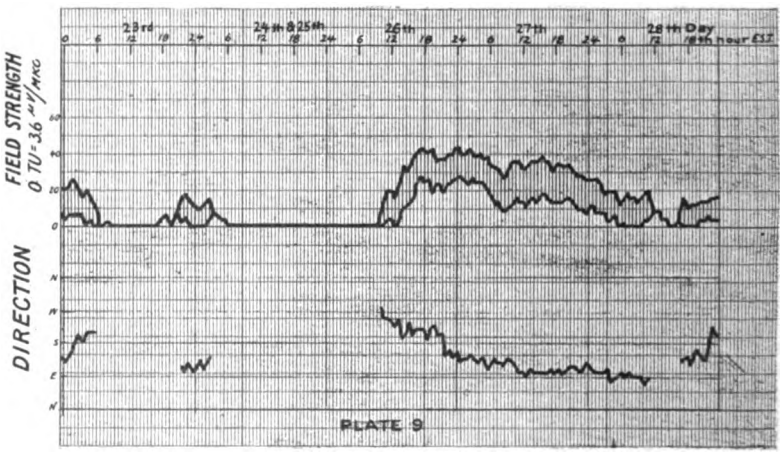

Fig. 9

The motor was so geared as to be able to make one revolution in six to ten minutes. From this it can be seen that each of the static graphs contains from three to ten complete revolutions of the compass and consumes from twenty minutes to 1½ hours.

The recorder used at Cliffwood by the Bell Telephone Laboratories is shown in the schematic diagram Fig. 10. It is an ordinary double detection set that requires altogether ten tubes, of which the last low frequency amplifier tube must be able to handle 10 watts in order to prevent overloading. The power supply may be rectified AC.

The gain control is inserted in the first intermediate frequency amplifier in order to be sure that no tubes are over-loaded. The

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local oscillator shown is used for amplification calibration of the set and requires no special shielding as its input voltage induced into the loop is comparatively large.

The selectivity of the set is determined by three separate units, viz., the antenna circuits, the intermediate frequency filter, and the low frequency filter, each of which has a specific use. Carson and Zobel[1] have made the following statement:

"In filters designed to select a band of frequencies of width W, the ratio of energy transmitted through the network by the signal and by random interference is inversely proportional to the band width and increases inappreciably when the number of sections is increased beyond two."

The main purpose of the filters is therefore not to define the frequency band of the set insofar as static is concerned, but to exclude continuous wave interference. It is hoped that 500 cycles

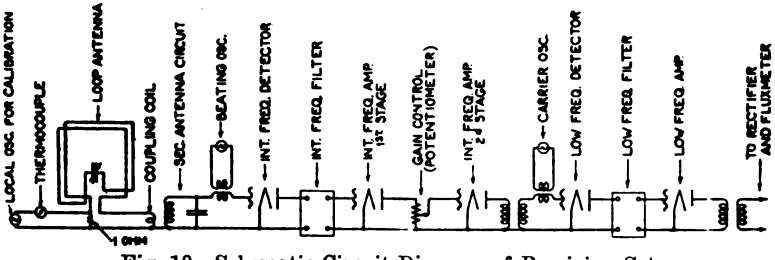

Fig. 10—Schematic Circuit Diagram of Receiving Set.

wide frequency bands[2] can be maintained free of c.w. interference for static measurements and the simplest way to obtain such a band in the receiver is to make the low frequency filter an efficient low pass filter that cuts off every frequency above 600 cycles. More than two coupled circuits are hardly required in the antenna circuits, but the intermediate frequency filter ought to have sharper cut-off points than two coupled circuits will give. The selection of filters naturally depends upon the c.w. interference and it may in some cases be possible to reduce the number of filters and thereby make the recorder cheaper. The records shown later correspond to a frequency band of 2000 cycles—between 57.5 and 59.5 kc.,—but it will probably not be long before c.w. interference makes it necessary to reduce this band width. It is desirable to have a loud speaker connected to the output of the set and occasionally listen for c.w. interference.

[1] "Transient Oscillators in Electric Wave-Filters"—John R. Carson and Otto J. Zobel, Bell System Technical Journal, Vol. II, No. 3, p. 27.

[2] Bands at 15, 30, 60, 120—kilocycles would probably be satisfactory.

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The constant-output control apparatus is shown in Fig. 11. The fluxmeter is seen in the upper right corner. Full deflection corresponds to 2×10^{-4} coulomb. The needle is normally free to move except when the cam Z presses the needle down until its point touches the scale OS. The shaft carrying the cam Z and the disk N is rotated one complete turn in 15 seconds by a clock motor. The different elements are explained in the figure and the whole action may be understood by studying this carefully. However, it is probably worth while to go through a complete 15 second period and explain in detail the purpose of each part.

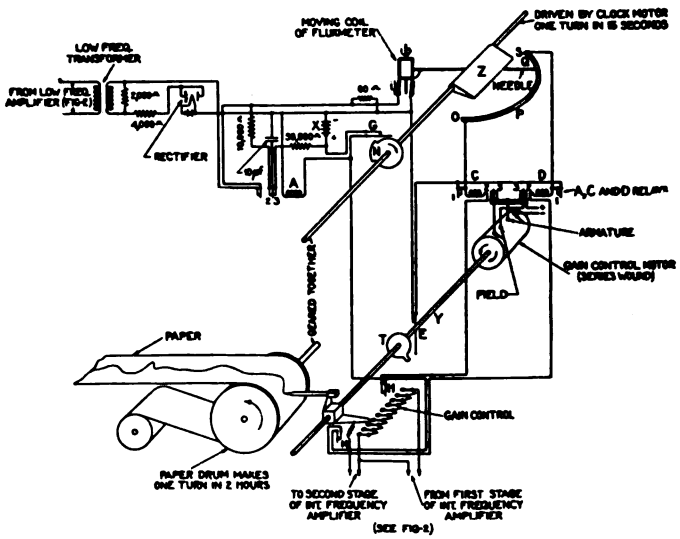

Fig. 11—Schematic Circuit of Constant-Output Control Apparatus.

Time in Seconds—0–10:

Switch G is open, therefore relay A is open and no current can pass through the windings of relays C and D. (These relays start the gain control motor, which is therefore shut off.)

Contact 1 of relay A is closed and closes the circuit consisting of the secondary winding of the low frequency output transformer, the rectifier for the static currents and the fluxmeter. The 2000- and 4000-ohm resistances in this circuit insure distortionless input voltage to the rectifier. The fluxmeter is damped by an 80-ohm shunt. The needle, which was initially at zero, will therefore move, its deflection being proportional to $\int i dt$.

Time in Seconds—10–14:

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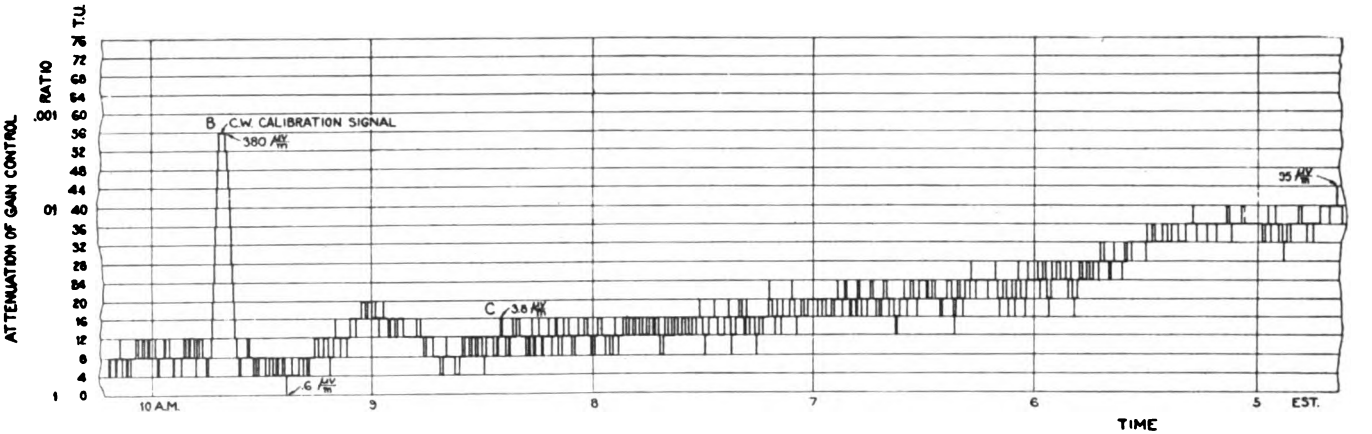

Fig. 12—Static Record, Morning of October 30, 1925, Cliffwood, N.J., U.S.A.

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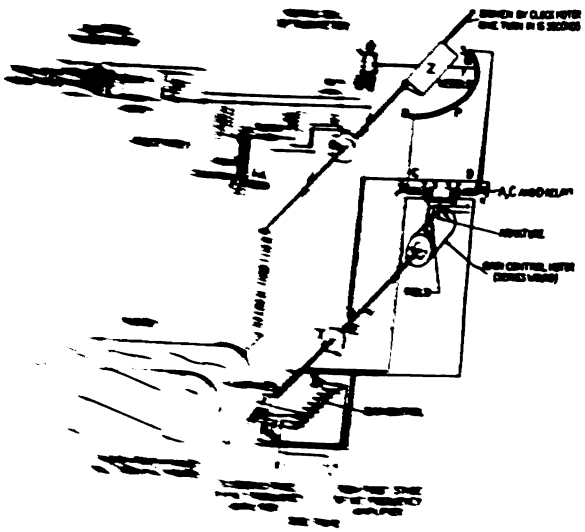
...apparatus is shown in Fig. 11. ... corner. Full deflection ... The needle is normally free to ... Z presses the needle down until its ... The shaft carrying the cam Z and the ... seconds by a clock motor. ... in the figure and the whole ... studying this carefully. However, ... complete 15 second ... each part.



...ing Control Apparatus.

... A is open and no current ... and D. (These relays ... shut off.)

... circuit consisting ... output transformer, ... ammeter. The 2000- ... distortionless input ... damped by an 80-ohm ... will therefore move,

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Switch G is closed by the cam on the revolving disk N and locks relay A.

Contact 1 of relay A is opened and opens the rectifier fluxmeter circuit, thereby bringing the fluxmeter needle to s stop.

Contact 3 of relay A is closed and makes the battery X charge the 10-μf. condenser through the 50,000-ohm resistance.

Time in Seconds—11–14:

The cam Z presses the needle point down on the scale OS. Now, one of three things will happen.

1. Static has decreased since the last period, so that the needle point will make contact with the metal strip OP and close the following circuit: Battery X, needle of fluxmeter, winding of relay C, switch H and switch G to battery X. Relay C is therefore closed and its closed contact 2, together with contact 3 of the open relay D will start the gain control motor. After approximately half a turn of the gain control or motor shaft Y the needle point is

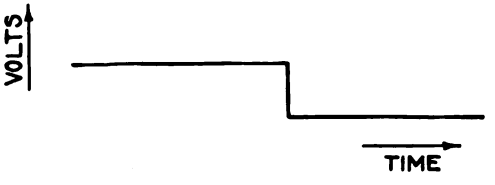

Fig. 13—Shape of Impulse Voltage.

lifted from OP by the rotation of the cam Z, but relay C stays closed due to the fact that it is self-locking through its contact 1, so that the shaft Y continues turning until the switch E is opened by the disk T. This opens the self-locking circuit of relay C. Relay C therefore opens and the gain control motor stops after the shaft Y has made exactly one complete turn and increased the gain of the set one step (4 Transmission Units or 1.58 times). Notice that the opening of the needle point contact does not break any current, due to the use of self-locking relays. This preserves the needle point contact.

2. Static has not changed since the last period. The needle point will now touch the insulating strip PQ and nothing else will happen, i.e., the gain of the set remains unchanged.

3. Static has increased since the last period so that the needle point now will make contact with the metal strip QS and close relay D and as in case 1 the motor will start and turn the shaft Y one turn, but this time in the opposite direction, i.e., the gain of the set is decreased one step.

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Time in Seconds—14:

Switch G is opened again by the revolving disk N and opens relay A. Contact 2 of relay A is closed and will discharge the 10-μf. condenser through the fluxmeter, thereby bringing the needle back to zero. (Notice that the time constant of this discharge circuit is $10{,}000 \times 10 \times 10^{6} = 1\text{–}10$ seconds.)

Time in Seconds—15:

A new period has started.

The purpose of the switches M and H is to stop the motor when the gain control switch arm has reached the end of the scale.

The recorder is of such recent development that no comprehensive data are yet available.

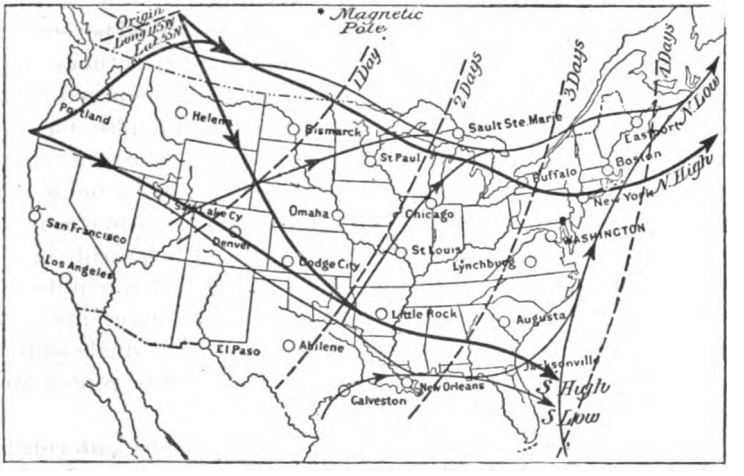

Fig. 14—The Bigelow System of Storm Tracks Across the United States.

Fig. 12 shows part of an actual record of static received on a set tuned to 57.5–59.5 kc. The ordinates represent the attenuation of the gain control of the set and it is to be remembered that the gain of the rest of the set is constant. The curve shows that the static power on the morning of October 30 changed more than 10,000 times. The point B on the curve gives the effect of inducing a local signal of strength 380 microvolts/m. in the loop.[3] The point C on the curve shows that at 8:25 A.M. the static intensity received on a 2000-cycle wide frequency band corresponded to the energy received from a c.w. signal of strength 3.8 microvolts/m.

[3] It may be worth while to have such a calibration signal introduced automatically for instance once every two hours.

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It would be practical always to relate static to such a c.w. signal. Experiments are now being conducted to determine whether the energy received from static is proportional to the width of the frequency band of the receiving set, and if such is found to be the case then it is proposed to have the data relate to a 1000-cycle wide band. That static is, say, 7 microvolts per meter per kilocycle (7 microvolts/m./kc.) would then mean that the energy of the static received on a 1000-cycle wide frequency band is the same as the energy received from a c.w. signal of strength 7 microvolts/m.

Attempts have been made to calibrate the set by inducing in the loop, voltages of the shape shown in Fig. 13. Relating static to such signals would have the advantage of being independent of the band width of the set. Such signals were obtained by closing and opening a mercury switch, but one signal per second, or 10 impulses per period, would overload the set (the tubes) very much. At least 10 impulses per second would be required if the set should not be overloaded by each individual impulse, but this would be a difficult task to accomplish and it is therefore recommended that static be measured as explained above, by inducing a local c.w. signal into the loop. The fact that five static crashes in the course of 10 seconds—one period—does not overload the set while 100 impulses of the shape shown in Fig. 5 are required to prevent overloading gives us some interesting information on static. It shows that a single static crash is not a single sudden change of the field in the ether and that it can not be represented by less than 20 consecutive impulses.

The record of Fig. 12 shows that each step on the gain control potentiometer is 4 *TU* and the selection of such steps and of 15 seconds will not be discussed. To decrease the 4 *TU* step to a 1 *TU* step would decrease the speed of the set, i.e., it would take four times longer for the recorder to register a sudden change in the static level which is particularly a disadvantage when the recorder is connected to a rotating directional antenna.

On the other hand a step larger than 4 *TU* would not give the static level with sufficient accuracy. If the time periods are changed from 15 to 10 seconds, then the "speed" of the set is increased, but the set is then inoperative over a larger part of the period since it takes 5 seconds to change the gain of the set and bring the fluxmeter needle back to zero. Besides, such a decrease in time period would increase the probability of overloading and also it would make the energy received per period

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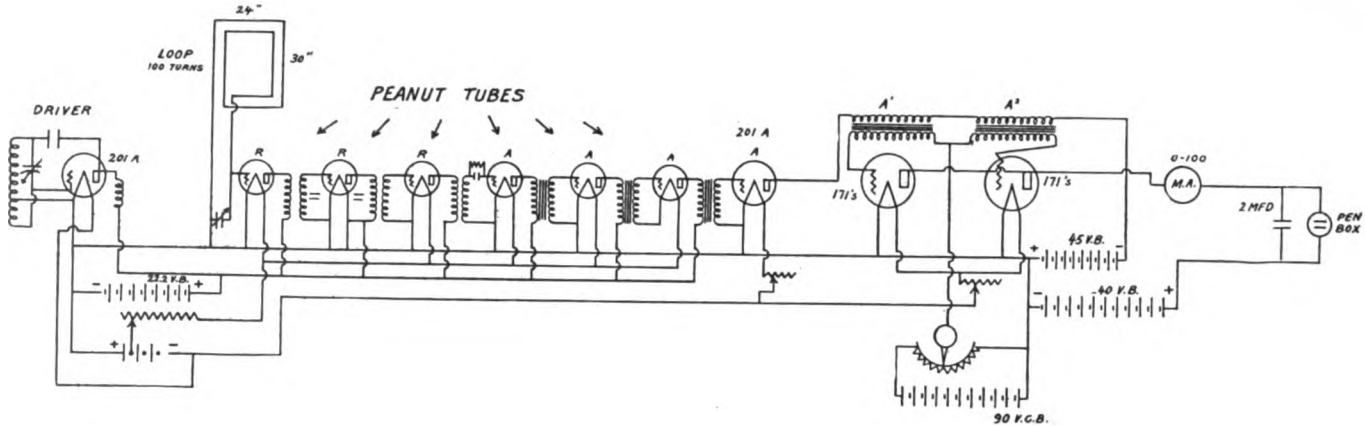

Fig. 15—Schematic Diagram of Kincaid Static Recorder. *Kittery* Installation.

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vary more irregularly especially if static consisted of separate crashes.

TABLE OF CONVERSION OF TRANSMISSION UNITS TO RATIOS

TU	Ratio
0	1
6	2
12	4
20	10
40	100
60	1000

Dr. Friis in describing his static recorder says, "The reason for the small advance which has been made to date in the automatic recording of static is probably due largely to the lack of suitable apparatus. Certainly there has never been any doubt that automatic records would be very valuable. It is just as important to know the static level as it is to know the strength of a radio signal because it is the static to signal ratio that determines the intelligibility of the signal. A static recorder connected to a rotating directional antenna system would tell us where static comes from and, therefore, enable the radio engineer to determine whether it is worth while to construct a directive antenna system. Also the connection between thunderstorm areas and static would make static recording valuable to the meteorological service. There is perhaps no reason why a suitable static recorder should not make it possible in a few years to obtain a daily static forecast just as we get our weather forecast now."

I believe another year will see such static forecast as Dr. Friis mentions possible, and with such forecasts will come a more exact knowledge of the relation of storms to static. The result of this will be the identification and tracking of many storms which are difficult especially over ocean areas where barometer readings are scarce, and the definite forecasting of communication conditions.

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A RADIO INTER-COMMUNICATING SYSTEM FOR RAILROAD TRAIN SERVICE

By

Henry C. Forbes

(Formerly of the Zenith Radio Corporation, Chicago)

1. Problem

In present day practice, communication between the front and rear ends of moving freight trains is carried on almost entirely by visual (arm or lantern) and whistle signals. Under certain conditions the air-brake pressure line may be used to transmit signals from rear to front of the train. Economic as well as safety considerations demand that some sort of communication be maintained between the conductor in the caboose and the engineman in the locomotive, principally in order to expedite the handling of the train.

Something of the difficulties encountered in the operation of trains of one hundred or more cars (approximately one mile) in length may be realized when it is considered that when using a visual communicating system, a heavy rain or snow storm, a heavy fog, curves in the track, hills, cuts, smoke, running "into the sun," and occasionally, diverted attention on the part of the engine crew, may partially or wholly prevent such communication. On occasions when stopping the train is imperative and the attention of the engine crew cannot be obtained by signaling from the caboose, it is possible to release the air at a valve provided for the purpose in the caboose, thereby setting the brakes at the rear end of the train and, if the train is not too long, reducing the air pressure in the locomotive so as to give warning that it is desired to pass a signal or to stop the train. On extremely long trains it is sometimes practically impossible to reduce the pressure in the locomotive sufficiently by this method to produce a definite signal. In such cases, the brakes are, of course, heavily set at the rear end of the train, and it is not uncommon to break a train in two or to pull it off the track while on a curve, in attempting to signal the locomotive in this manner.

At the instigation of Committee No. 12 (Radio and Wire Carrier Systems) of the American Railway Association, the Zenith

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Radio Corporation undertook, in January, 1926, the development of a radio inter-communication system for such freight service. In response to a questionnaire submitted to that committee, it was ascertained that it was the opinion of the interested railroad that while either a one-way or two-way radio signaling system would be a distinct improvement over the existing methods of signaling, what is really desired is a complete two-way telephonic inter-communicating system, providing, of course, that such a system can be made sufficiently strong and simple of operation to be entirely reliable. The New York Central Railroad, through the office of the General Superintendent of Telegraph and Telephone, and Committee No. 12 of the A.R.A., volunteered the use

Fig. 1—Showing Length of Train Equipped With Radio Intercommunicating System.

of its equipment and facilities for any tests or experiments which might be deemed necessary during the development of the apparatus. During the subsequent development of the apparatus, several conferences with members of a subcommittee appointed for the purpose were held in order to correlate, in so far as possible, the development of the apparatus with existing standard railroad practice.

2. Requirements of the Service

Several rather difficult problems were presented in the design of suitable equipment for this service. There is no source of electric power on a caboose, and it was not considered practicable to utilize

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the power from the turbine-driven headlight generator on the locomotive. Battery-operated sets were therefore necessary at each end of the train. It was necessary to provide sufficient power to produce an understandable loudspeaker signal at the opposite end of the longest train that might be encountered, while the train was running at top speed. At the same time it was desirable to limit the power used in order to limit the size of the apparatus and the drain on the batteries, and also to minimize the possibility of interference should the use of such apparatus become general.

The choice of a suitable antenna for the locomotive was a considerable problem. Modern road engines are constructed entirely of steel with an overhead clearance of about five inches at the center of the cab roof, and only about 20 inches at either side of the cab. It was desirable that the antenna used should be mounted entirely on the locomotive, so that the coaling operations and the taking of water would not be interfered with. The antenna must be constructed within the overhead clearance limits, must be sufficiently sturdy to withstand road service, and should transmit efficiently under all weather conditions, and in all directions in the horizontal plane so that the signals will not fade while making sharp curves.

The mounting of the radio apparatus itself required some attention inasmuch as it is necessary that it be installed in an extremely small space. It must be completely shielded, should withstand water and soot and exposure to heat, and must be extremely carefully cushioned against the terrific jarring and battering to which it is continually subjected in railroad service. Sturdiness and simplicity of operation are essential.

3. Development

Preliminary tests of the apparatus undergoing development were made between a fixed station at the Zenith laboratory, constructed to simulate the installation on a train, and a mobile station installed in an automobile. Several tests and experiments were also made in the Chicago yards of the New York Central Railroad to determine the characteristics of transmission from antennas installed on the road engines. The form of antenna finally selected consists of a single-turn horizontal loop mounted closely above the cab roof on the locomotive and around the cupola on the caboose. Such an antenna is nearly non-directional in the horizontal plane parallel to the ground, and the transmission may therefore take place in any direction from either end of the train.

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Experiments have shown that the range of low-power transmitters using this form of antenna on a locomotive is limited to something under ten miles when transmitting to a similar form of antenna, although very satisfactory communication is obtained at lesser distances. This is a distinct advantage in railroad service inasmuch as it materially lessens the interference problem where a number of such installations are to be made in a given area. It has also been found that the strength of the received signal and the range of the apparatus may be improved by making use of the vertical as well as the horizontal components of the transmission at the receiving end. Such attempts, however, are likely to lead to pronounced directional effects. Only the horizontal components of the radiated energy were utilized in these experiments. The loop antennas used were constructed of brass piping and fittings mounted on glazed porcelain high-voltage pin insulators. The insulator brackets were bolted directly to the cab of the locomotive and to wooden supports on the caboose. Glazed porcelain entrance bushings were used in the roof of both cab and caboose.

The Department of Commerce has assigned a band of frequencies between 2300 and 2750 kilocycles (109 and 130 meters in wavelength) for mobile services. A frequency of 2600 kilocycles (wavelength of 115.4 meters) was adopted for use with this apparatus, this frequency being particularly well adapted to the single-turn loop antennas.

Duplicate two-way telephonic transmitter-receivers were mounted (for the purposes of demonstration) in an iron box 30 in. long, 6 in. deep, and 10 in. high, the receiving apparatus occupying a left-hand compartment and the oscillator circuits a right-hand compartment, with the modulator, speech amplifier and switching relay circuits in the center. A simple regenerative receiving circuit was employed having a detector and three stages of audio-frequency amplification. A power tube was used in the third stage. A single tuning control, with fixed regeneration is thereby possible, and has been found adequate.

The transmitter employed two tubes of 50-watt rated plate dissipation, one as oscillator and the other as modulator, with a $7\frac{1}{2}$-watt tube as a speech amplifier. Power for the plate circuits of the transmitting tubes was obtained from an Electric Specialty Company dynamotor of 200-watt capacity mounted in a totally enclosed frame, and provided with ball bearings. This dynamotor starts and stops very quickly. The dynamotor is driven by a

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twelve-volt Exide storage battery of the heavy-duty vehicle type, although the system is readily adaptable to any battery voltage. Power for lighting all filaments is obtained directly from this same battery.

The filter system for the dynamotor and the starting relay are mounted together in a heavy box arranged with suitable terminals. Another box houses the necessary dry batteries for supplying the plate circuits of the receiving tubes. The life of the tubes should be about 1000 hours, and the dry batteries must be replaced at approximately monthly intervals. Storage batteries

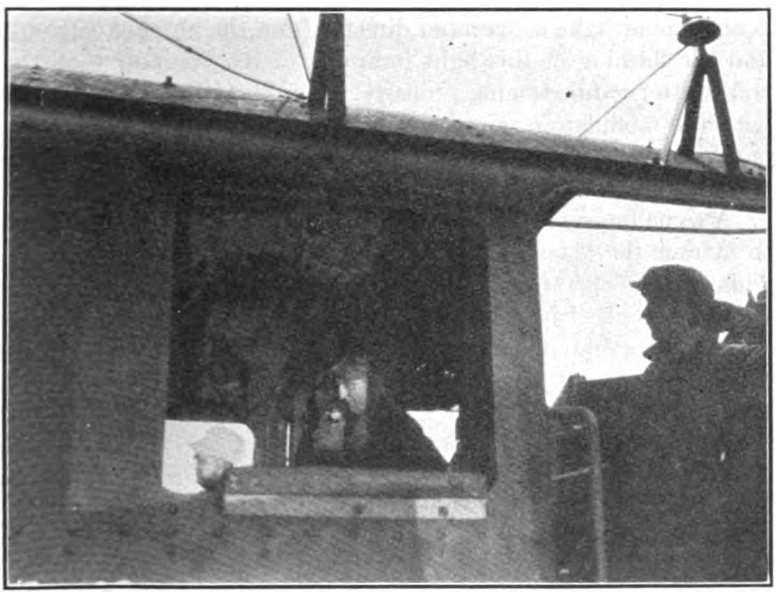

Fig. 2—Transmitter-Receiver Installed in Locomotive Cab, With Fireman Operating.

of 175 ampere-hour capacity were used, this being sufficient to operate the circuits under ordinary conditions for at least two trips across a division. All switching operations are done by means of relays actuated by a "press to talk" button on the handle of a "grab-o-phone" type of hand microphone on which a telephone receiver is also provided. On the experimental installations of this apparatus, both the loudspeaker and telephone receiver operate simultaneously.

In practice, the receiving sets are left in operation at all times when the train is under way, a switch being provided for the

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purpose. When communication is desired, the button on the handle of the microphone is pressed, thereby actuating the relays which make the necessary switching connections for operating the transmitter. The button is released to restore the receiver to operation. The switching, including the starting of the dynamotor, takes only about one second, and very rapid two-way conversation is possible.

The functioning of the apparatus is indicated to the operator by the flashing of colored pilot lights. One light indicates that the receiver is in operation. A second flashes when the "talk" button is pressed, and indicates that the transmitter is ready for operation. A third pilot light is operated directly from the antenna current, and the flashing of this light indicates to the operator that the transmitter is functioning properly. This third light also flickers with the modulation, giving a direct indication that modulation is taking place. The meters shown in the photographs are not to be used on permanent apparatus.

A signaling system has also been provided which may be used to attract the attention of the crew at either end of the train. This signal is operated by pressing a second "signal" button which is mounted either on the microphone handle or on the set proper, and, when operated, produces a shrill note of about 800 cycles in the loudspeaker at the opposite end of the train. This frequency was selected for this purpose after some experiment, and is readily discernible in the locomotive through the noises encountered in running. This signal may also be used to pass code signals in case of failure of the telephonic system.

Special microphones were used in order to avoid the introduction of the terrific road noises into the communicating system. An aircraft type of anti-noise microphone, originally developed for war service, was found to be very satisfactory.

Extraordinary cushioning of all apparatus in both transmitter and receiver was necessary in order to protect the parts properly from the destructive vibration and shock encountered in railroad work. Combinations of heavy sponge rubber and spring suspensions were found to be generally satisfactory, although some difficulty has been experienced in sufficiently quieting the detector tube.

For the purposes of demonstration, one set was mounted on the inside of the rear bulkhead of the locomotive cab over the rear opening. Special plugs and cordage connected the set with the batteries, dynamotor, and filter, which were mounted in a shovel-

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rack just outside of this same bulkhead. The leads to the loop antenna were run directly up through the cab roof. A loudspeaker horn was mounted just to the rear and above the head of the engineer, so that the sound was projected directly past his head. The hand microphone, with its control button was placed in a clip convenient to both the engineer and the fireman.

In the caboose the set was installed flush into the side of a locker convenient to the conductor's desk, with the batteries, dynamotor, and filter mounted on the floor of the locker. The leads to the loop antenna went out directly through the roof. The loudspeaker in this case was mounted overhead at one end of the car.

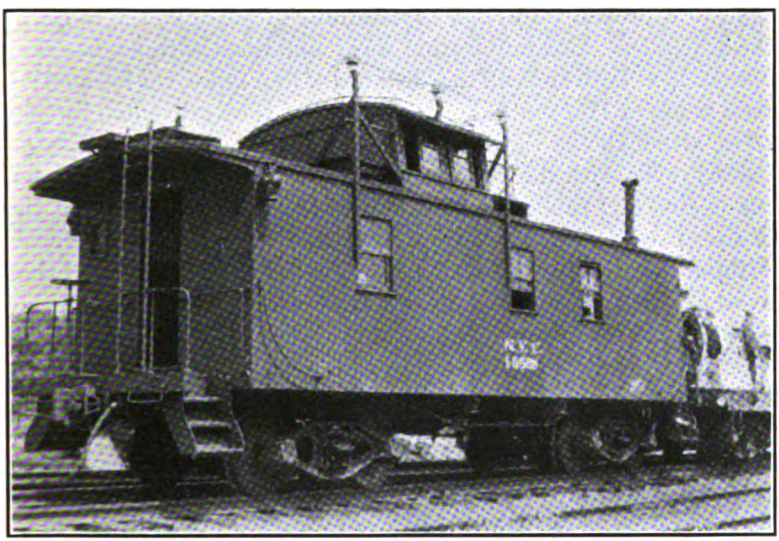

Fig. 3—Antenna on Caboose.

4. Preliminary Tests

In the first of the preliminary road tests of the apparatus on June 11, 1926, a special train of twelve cars was run over the western division of the New York Central between Elkhart, Ind. and Chicago (Englewood), a distance of about 100 miles. The apparatus operated in a manner considered very satisfactory by the several railroad men present. No difficulty was had in maintaining two-way conversation between the caboose and the locomotive when standing still or when running at top speed. The engineer was able to hear and understand everything coming from

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the loudspeaker behind his head without diverting his attention from the road ahead. The fireman was able to handle the apparatus and to carry on extensive conversations with the officials in the caboose. All train orders were given by means of the radio installation, and those present, including the train crew, were quite enthusiastic as to the possibilities of such radio apparatus. A severe rain and electrical storm was encountered over practically the entire run, which occupied about five hours, and aside from the sharp crashes heard in the loudspeakers, it caused no trouble whatsoever. Near Pinola, Ind., the train was stopped and the engine uncoupled from the train and run ahead under orders given entirely by radio from the caboose, in order to simulate the conditions of a broken train, and to determine if possible, the range of the apparatus. Communication was maintained with loudspeaker operation at both ends up to a distance of four miles, the locomotive then being at Hudson Lake, Ind. Lack of time prevented further tests, although the signals at that distance were still quite satisfactory, even with the locomotive in motion. The country between the engine and caboose during this test was rather hilly, and the engine passed through several cuts and around several curves without affecting the operation. During this test it was found that overhead power wires and small signal bridges had practically no effect on the operation of the system. When passing under a large steel bridge, however, communication is broken (except for weak headphone signals) while either the locomotive or caboose is actually under the bridge. The communication is not affected by the presence of the bridge between the engine and caboose after the engine has once cleared it. It was also found that there were no interfering disturbances produced on adjacent land telephone wires alongside the right-of-way, even with the caboose transmitting within 30 feet of a phone box. The passing of a train going in the same or opposite direction upon an adjacent track was found to have no effect upon the operation.

Subsequent tests in the Chicago yards have indicated that satisfactory communication may be maintained between the caboose and locomotive up to a distance of six miles with this apparatus. Beyond this distance only headphone signals are possible. Communication has actually been held at a separation of eight miles between a moving locomotive and a stationary caboose. It is expected that tunnels will completely prevent communication while either end of a train is in or near the tunnel.

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5. Official Tests

An official demonstration of the apparatus before Committee No. 12 of the American Railway Association and a large party of officials of various railroads and representatives of the press was given on July 8, 1926, on a New York Central freight train operated between Chicago (Englewood) and Elkhart, Ind. There were 116 cars in the train, of which approximately sixty were loaded. A coach and a business car were placed before the caboose for the accommodation of the representatives. During the run of approximately five hours the practicability of a radio communication service of this nature was clearly demonstrated. A number of orders were transmitted which saved the delays which would have otherwise occurred while a member of the train crew walked the entire length of the train. Practically all of the representatives carried on a conversation with the engine crew at some time during the run. With the exception of minor difficulties caused by the loosening of cordage connections by the severe vibration, no particular difficulty was encountered in maintaining complete and satisfactory communication between the caboose and the engine at all speeds. During the run, the train was double-headed and a pusher locomotive was used directly behind the caboose for a short distance. This apparently had no effect upon the operation of the radio system. The consensus of opinion of those present was undoubtedly that the application of radio communicating systems to railroad service was in immediate prospect.

In future designs of apparatus for this purpose, it is proposed to provide several fixed wavelength adjustments on the transmitters, any one of which may be selected at will by the manipulation of a switch. It is contemplated that instructions will be given to the train crew before the run is started as to which of these several wavelengths is to be used by that train during its run. Trains traveling in the opposite direction would be operated upon another wavelength and there would be no interference between the trains while they were passing or within range. Communication between the two trains would be readily possible at any time they were within range, however, by merely turning the master switch to the wavelength of the other train and proceeding in the usual manner. The receivers would be made with a single wavelength control adjustment which would cover all of the wavelengths in use. Inasmuch as there are a number of channels available within

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the band assigned to this service, there should be no difficulty with interference for some time to come.

6. Other Uses

With the application of radio communication to railroad freight service comes also the application of a similar service to passenger train operation, not only for the purpose of train handling, but also for the purpose of intercommunication between trains, and between a train and a way-station. A further application which has been given some consideration is that to large freight yards where one towerman controls all of the locomotives working in a yard perhaps five miles in length. Much time is now lost because of the lack of prompt communication between the tower and the individual locomotives.

Summary

The problem of communication between the front and rear ends of long freight trains is briefly discussed, and a two-way telephonic radio intercommunicating system developed for this service is described. The results of tests of the apparatus under working conditions are also given.

Acknowledgment

The writer wishes to acknowledge sincerely the hearty cooperation and assistance of the officials of the American Railway Association and the New York Central Railroad in the conduct of the experimental work during the development of this apparatus. Particular mention is also due H. H. Meinhard and G. Gustafson of the Engineering Department of the Zenith Radio Corporation for their part in the development.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued August 30, to September 13, 1927

By

JOHN B. BRADY

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,640,393—OPTICAL MEANS FOR GENERATING, AMPLIFYING, AND CONTROLLING ELECTRIC CURRENTS—WILLIAM W. COBLENTZ, of Washington, D. C. Filed Sept. 18, 1923, issued Aug. 30, 1927.

1,640,427—RADIO RECEIVING SYSTEM—C. W. RICE, of Schenectady, N. Y. Filed Nov. 21, 1922, issued Aug. 30, 1927. Assigned to General Electric Co.

1,640,462—ELECTRON DISCHARGE DEVICE—R. C. MATHES, of Wyoming, N. J. Filed Oct. 16, 1924, issued Aug. 30, 1927. Assigned to Western Electric Co., Inc.

1,640,503—ELECTRON DISCHARGE DEVICE—W. G. HOUSKEEPER, of New York, N. Y. Filed Oct. 30, 1922, issued Aug. 30, 1927. Assigned to Western Electric Co., Inc.

1,640,534—RADIO ANTENNA SYSTEM—F. CONRAD, of Pittsburgh, Pa. Filed Dec. 14, 1920, issued Aug. 30, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,640,561—VARIABLE CONDENSER—E. W. BREISCH, et al, of Edgewood Park, and LAWRENCE R. GOLLADAY, of East Pittsburgh, Pa. Filed July 24, 1925, issued Aug. 30, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,640,710—ELECTRON EMITTING CATHODE AND PROCESS OF PREPARING THE SAME—H. Miller, of Newark, New Jersey. Filed March 1, 1926, issued Aug. 30, 1927. Assigned to Hazeltine Corp.

1,640,929—PORTABLE RADIO ANTENNA—H. FARKOUH, of Brooklyn, N. Y. Filed Nov. 24, 1925, issued Aug. 30, 1927.

1,641,289—RADIO SIGNALING SYSTEM—RUSSELL S. OHL, New York City. Filed Nov. 23, 1925, issued Sept. 6, 1927. Assigned to American Telephone and Telegraph Co.

1,641,314—VARIABLE CONDENSER—HARRY L. BRADLEY, Milwaukee, Wisconsin. Filed Oct. 17, 1924, issued Sept. 6, 1927.

1,641,389—CONDENSER WINDING MACHINE—F. C. KROEGER and HARRY E. NORVIEL, Anderson, Ind. Filed July 29, 1922, issued Sept. 6, 1927. Assigned to Delco-Remy Corp.

1,641,395—RECTIFYING RADIO SHIELD—TOM MOORE, Cincinnati, Ohio. Filed Sept. 15, 1924, issued Sept. 6, 1927.

1,641,425—TRANSMISSION CONTROL—FRANK H. GRAHAM, Brooklyn, N. Y. Filed Oct. 24, 1923, issued Sept. 6, 1927. Assigned to Western Electric Co.

1,641,431—COMMUNICATION SYSTEM—JOSEPH W. HORTON, Bloomfield, N. J. Filed Dec. 15, 1925, issued Sept. 6, 1927. Assigned to Western Electric Co.

1,641,438—VARIABLE CONDENSER—L. L. JONES, Oradell, N. J. Filed Dec. 13, 1923, issued Sept. 6, 1927.

1,641,513—SUPPORTING MEANS FOR ELECTRICAL DEVICES—J. A. WARREN. Filed May 6, 1925, issued Sept. 6, 1927. Assigned to Dubilier Condenser Corp. (J. A. Warren, Mount Vernon, N. Y.)

1,641,593—MULTIPLE VARIABLE CONDENSER—FREDERICK A. KOLSTER, Palo Alto, Calif. Filed Apr. 28, 1926, issued Sept. 6, 1927. Assigned to Federal Telegraph Co.

1,641,608—ELECTRICAL SIGNALING—G. O. SQUIER, Washington D. C.; J. O. MAUBORGNE, Chicago, Ill.; and LOUIS COHEN, Washington, D. C. Filed June 24, 1922, issued Sept. 6, 1927.

1,641,635—ADJUSTABLE CONDENSER—R. M. KLEIN, New York City. Filed Nov. 1, 1924, issued Sept. 6, 1927. Assigned to F. A. D. Andrea, Inc.

1,641,687—VACUUM CONDENSER—H. J. NOLTE, Schenectady, N. Y. Filed Oct. 8, 1925, issued Sept. 6, 1927. Assigned to General Electric Co.

1,642,173—RADIO SIGNALING SYSTEM—H. J. ROUND, London, England. Filed Mar. 2, 1922, issued Sept. 13, 1927. Assigned to Radio Corp. of America.

1,642,350—TELESCOPING RADIO LOOP OR ANTENNA—L. ADAMS, Buffalo, N. Y. Filed Feb. 6, 1926, issued Sept. 13, 1927.

1,642,420—RADIO RECEIVING APPARATUS—S. LOEWE, Berlin, Germany. Filed Mar. 18, 1927, issued Sept. 13, 1927. Assigned to Radio Corp. of America.

1,642,498—ELECTRON DISCHARGE DEVICE—W. G. HOUSKEEPER, New York, N. Y. Filed Oct. 30, 1922, issued Sept. 13, 1927. Assigned to Western Electric Co.

1,642,522—LOOP ANTENNA—E. VROOM, Ossining, N. Y. Filed May 26, 1924, issued Sept. 13, 1927. Assigned to Western Electric Co.

1,642,526—MEANS FOR SUPPLYING POWER TO THERMIONIC VALVES—G. M. WRIGHT, Chesterfield, England. Filed Aug. 17, 1926, issued Sept. 13, 1927. Assigned to Radio Corp. of America.

1,642,663—SYSTEM OF RADIO COMMUNICATION—E. L. CHAFFEE, Belmont, Mass. Filed Aug. 11, 1922, issued Sept. 13, 1927. Assigned to John Hays Hammond, Jr.

1,641,945—ELECTRICAL INSTRUMENT—H. J. MURRAY, Brooklyn, N. Y. Filed Nov. 6, 1923, issued Sept. 6, 1927.

1,641,946—MOVABLE DIELECTRIC CONDENSER—H. J. MURRAY, Brooklyn, N. Y. Filed Jan. 29, 1925, issued Sept. 6, 1927.

1,642,011—LIGHT TELEPHONY—L. W. CHUBB, Edgewood Park, Pa. Filed June 15, 1921, issued Sept. 13, 1927. Assigned to Westinghouse Electric & Mfg. Co.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED SEPTEMBER 7, 1927

Transferred to the Fellow Grade

New Jersey,	Upper Montclair	Hallborg, H. E.
Louisiana,	New Orleans	Lehde, P. E.

Transferred to the Member grade

Massachusetts,	Boston	Harris, C. C.
New York,	Newburgh	Thomas, W. A.
	New York	Amy, E. V.
	New York	Herbert, Harold
Australia,	Melbourne	Hull, Ross A.

Elected to the Member grade

Dist. Columbia,	Bellevue, Anacostia	Owens, R. B.
New Jersey,	Mountain Lakes	Asserson, Raymond
Australia,	Melbourne	Bearup, T. W.
England,	Liverpool	Bailey, W. H.
Scotland,	Greenock	Murchie, John

Elected to the Associate grade

California,	Livermore, USVB Hospital 102	Ryan, R.
	Oakland, Hotel Oakland	Allen, Preston D.
	San Francisco, 866 Post Street	Lambert, R. H.
	San Francisco, 1008 Larkin Street	Powell, L. H.
Georgia,	Atlanta, 82 Woodland Ave.	Howard, R. E.
	Atlanta, 594 Luckie St., N. W.	Jones, James M., Jr.
	Atlanta, 1513 S. Gordon St., N. W.	Walker, R. L.
	Columbus, 1712 Wynnton Drive	Brown, J. L.
	Columbus, Levy-Morton Elec. Co.	Cody, W. G.
Illinois	Chicago, 738 Hearst Square	Hurja, A. O.
	Chicago, 1714 Foster Ave.	Walgren, C. A.
Indiana,	Loganport, 514 Broadway	Vivian, J. A.
	Valparaiso, P. O. Box 225	Lucas, S. M.
Iowa,	Davenport, care Station WOC	Elliott, F. W.
	Iowa City, 728 Bowery St.	O'Brien, E. J.
Kansas,	Topeka, S. W. Bell Tel. Co.	Gordon, H. C.
Louisiana,	New Orleans, care Caramel Fruit Co.	Carleton, Thos.
Maryland,	Baltimore, 4205 Oakford Street	Ranft, W. Q.
Massachusetts,	Boston, 1111 Commonwealth Ave.	Abbott, R. F.
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	Somerville, 21 Eastman Road	Trefry, L. D.
	Worcester, 21 Hackfeld Road	Adams, Stuart
Montana,	Havre, Box 957	Johnson, C. A.
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	Trenton, 152 South Olden Avenue	Kale, S. S.
	Woodbridge, 118 Schoder Ave.	Tappen, G. L.
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	Brooklyn, 348 Douglass St.	Larson, J. M.
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	179 Clapham Road, London SW 9,	Webb, J. H.
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Mexico,	Benito Juarez 39, Vera Cruz,	Angel Fernandez, M.
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WARREN FETTER HUBLEY
Late Treasurer, Institute of Radio Engineers, 1915-1927

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 15 November, 1927 Number 11

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GENERAL INFORMATION

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INSTITUTE ACTIVITIES

October Meeting of the Board of Direction

At the meeting of the Board of Direction of the Institute, held on October 5, 1927 in the Institute Offices, the following were present: Dr. Ralph Bown, President; Frank Conrad, Vice-President; Dr. A. N. Goldsmith, Secretary; L. A. Hazeltine, R. A. Heising, J. V. L. Hogan, R. H. Marriott and J. M. Clayton, Assistant Secretary.

Mr. Melville Eastham was elected Treasurer of the Institute to fill the unexpired term caused by the death of Mr. Warren F. Hubley.

Upon recommendation of the Committee on Admissions, the following applications for transfer or election to the higher grades of membership in the Institute were approved: Transfer to the grade of Fellow: M. C. Batsel and L. E. Whittemore; transfer to the grade of Member: Carlton D. Morris and Conway Taylor; election to the grade of Member: John F. Grinan, E. B. Moullin and G. McB. Salt.

One hundred and two Associate and ten Junior members were elected.

In the absence of nominations by petitions, the ballot for 1928 officers of the Institute will be as follows:

For President, Professor E. M. Bennett and Dr. A. N. Goldsmith.

For Vice-President, A. H. Grebe and L. E. Whittemore.

For Managers, Dr. W. G. Cady, Dr. J. H. Dellinger, W. G. H. Finch, and R. H. Manson.

Liebmann Memorial Prize

The Liebmann Memorial Prize for this year was awarded to Dr. A. Hoyt Taylor for his work in connection with the investigation of radio transmission phenomena. The award will be made at one of the sessions of the 1928 convention of the Institute early in January.

1928 Year Book

With the ballots for 1928 Officers of the Institute, which will be in the mail shortly after November 1st, there will be found a form to be used by all members in notifying the Institute of any changes of business or residential address for

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inclusion in the 1928 Year Book. These forms should be executed and returned with the ballots. All members are urged to return both promptly in the envelope provided therefor.

The 1928 Year Book will contain a catalog of the membership of all grades, as of December 31, 1927. Changes in address or unpaid members cannot be included in the membership list if the Institute is not notified prior to that date.

Membership Cards

At the October meeting of the Board of Direction it was decided that membership cards for *all* grades of membership shall be available next year. These cards will be issued only upon request.

Institute Meetings

New York Meeting

At the New York meeting of the Institute, held on October 5, 1927 in the Engineering Societies Building, 33 West 39th Street, New York City, a paper on "Method of Reducing The Effect of Atmospheric Disturbances" was presented by Major E. H. Armstrong.

In the discussion which followed the paper, the following took part: Dr. A. N. Goldsmith, Carl T. Englund, Roy Weagant, R. H. Marriott, D. K. Martin, A. H. K. Morse, A. E. Reoch, and others.

The attendance at this meeting was over six hundred.

A limited number of copies of the paper, in pamphlet form, are available to members of the Institute upon application to the Institute Offices.

Atlanta Section

A meeting of the Atlanta Section of the Institute was held in the Atlanta Athletic Club on September 7, 1927. H. L. Wills, Vice-Chairman, presided. Mr. Oliver, of the Southern Bell Telephone Company, read the paper of Messrs. Diamond and Webb on "The Testing of Audio-Frequency Transformer-Coupled Amplifiers."

Following the paper and discussion, the motion picture "Silent Drama of Telephony" was shown.

The attendance at this meeting was twenty-five.

Buffalo-Niagara Section

At the invitation of the local Section of the American Institute of Electrical Engineers, the Buffalo-Niagara Section

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of the Institute of Radio Engineers attended an A.I.E.E. meeting on September 23, 1927 at which E. C. Markley of the General Electric Company delivered a paper on "Carrier Current Telephony." This paper was discussed by C. A. Boddie and others.

A meeting of the Buffalo-Niagara Section was held on October 12, 1927 in Townsend Hall, University of Buffalo. L. C. F. Horle was the presiding officer.

A paper by Francis H. Engel, of the Radio Corporation of America, on "Notes on A-C. Tubes" was presented by Mr. Engel. Messrs. Hector, Stone, Horle, Jones and others participated in the discussion which followed.

The attendance was fifty-five members.

The next meeting of the Section will be held on November 17, 1927 in the Science Hall, University of Buffalo, at which time Dr. Ralph Bown, President of the Institute, will present a paper on "Transatlantic Telephony."

Canadian Section

At the September 14th meeting of the Canadian Section, held in the Electrical Building of the University of Toronto, A. M. Patience presided.

C. I. Soucy, of the University of Toronto, presented the first of a series of "Junior Lectures" prepared for the benefit of the younger members of the Section. The first lecture was entitled "Inductance Coils." It is planned that at each meeting of the Section, for a period of about twenty minutes, one of these Junior Lectures will be given. The proposed subjects include Coils, Condensers, Application of Coils and Condensers, Tubes, Radio-Frequency Circuits, Audio-Frequency Circuits, Sensitivity and Selectivity, Loudspeakers, Batteries and Rectifiers, Installation and Service Problems and other subjects.

The paper of the evening, entitled "Rectox, The Oxide-copper Rectifier", was presented by A. L. Atherton. In the discussion which followed, the following took part: Messrs. C. I. Soucy, C. L. Richardson, D. Hepburn, J. Leslie, G. E. Pipe, C. A. Lowry, A. R. Starr and others.

Cleveland Section

On September 23, 1927 a meeting of the Cleveland Section, J. R. Martin presiding, was held in the Auditorium of the new School of Lighting, General Electric Company, Nela Park. Wm. T. L. Cogger, Manager of the Vacuum Tube Department

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of the Nela Works, presented a paper on "The New Vacuum Tube (A-C.)".

Seventy-five members of the Section attended the meeting.

On November 4, 1927, Dr. Ralph Bown, President of the Institute, will present a paper "Transatlantic Telephony" in the Case School of Applied Science, Cleveland.

Philadelphia Section

The September meeting of the Philadelphia Section was held on the twenty-third of the month in the Bartol Laboratories. J. C. Van Horn presided.

A paper by Dr. Ralph Bown on "Transatlantic Radio Telephony" was presented by Dr. Bown.

Messrs. Miller, Arnold, Snyder and others took part in the discussion which followed.

The attendance was fifty-two members.

San Francisco Section

The San Francisco Section is planning a renewal of its meetings beginning with a session in November, to continue throughout the winter and spring months.

At a luncheon in the Engineers' Club, San Francisco, on October 7, 1927 the affairs of the Section were discussed by Dr. L. F. Fuller (member of the Committee on Sections), D. B. McGown, B. H. Linden, A. Y. Yuel, and Past-President Donald McNicol.

The large growth of radio on the Pacific Coast now makes possible the presentation of a wide range of papers on radio and related subjects of interest to members of the Institute.

Seattle Section

On September 17, 1927 a meeting of the Seattle Section was held in the Telephone Building, 3rd and Senaca Streets, Seattle. Mr. T. M. Libby was Chairman.

Two papers were presented. The first, by J. R. Tolmie, was a discussion of the Preliminary Drafts of Reports of the Institute's Sub-committees on Vacuum Tubes, Receiving Sets and Electro-Acoustic Devices. The following participated in the discussion of these preliminary drafts: Messrs. Libby, Willson, Burleigh, and others.

The second paper, by Howard F. Mason, was entitled "Short Wave Transmitter Operation in the Arctic." Messrs. Libby, Burleigh, and Tolmie discussed this paper.

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Twenty-eight members of the Section were present at the meeting.

Proposed Sections

Continued correspondence is being held with members of the Institute looking to the formation of Sections of the Institute in the following localities: Milwaukee, New Orleans, Pittsburgh, Denver, and Minneapolis.

It is anticipated that Sections will be formed in several of these cities before the first of next year.

Committee Work

Sub-Committee of the Standardization Committee

A meeting of the Sub-committee on Determination of Circuit Constants was held at Institute Headquarters on September 23, 1927. The following members of the Sub-committee were present: H. M. Turner, Chairman; C. T. Burke, G. C. Crom, and G. S. Southworth.

Miscellaneous questions pertaining to symbols and nomenclature which have been referred to the Committee were discussed and for the most part agreed upon. Certain of these matters are to be taken up with the American Institute of Electrical Engineers with the hope of securing uniformity of action.

The greater portion of the day was devoted to a consideration of a partially completed preliminary draft of a report. One of the major problems is that of obtaining convenient and reliable methods of measuring the inductance of iron-cored coils with and without superposed direct current. The Chairman has spent considerable time studying the various methods in general use and devising others and comparing them. Some of this material was submitted to the Committee members present at this meeting for their comments and suggestions. Other problems discussed included determining the characteristic curves of audio-transformers, measuring the capacity of large and small condensers, the distributed capacity of coils, and measuring the resistance of grid leaks.

Committee on Admissions

At its meeting held on October 5, 1927, the following members of the Committee on Admissions were present: Vice-President, Frank Conrad, Chairman; Professor L. A. Hazeltine,

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Dr. L. M. Hull, F. H. Kroger, and H. F. Dart. The Committee acted upon eleven applications for transfer or election to the higher grades of membership in the Institute.

Committee on Membership

A meeting of the Committee on Membership was held in the Institute Headquarters on September 23, 1927. At this meeting H. F. Dart, Chairman, E. R. Shute, and J. M. Clayton were present.

The Committee has many plans before it looking to circularization and other activities which will result in an increase in the membership in the Institute.

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OBITUARY

With deepest regret The Institute of Radio Engineers announces the death of

Warren Fetter Hubley

Mr. Hubley's interest in the development of The Institute of Radio Engineers goes back to the early days of the organization. He first became interested in the meetings of the Wireless Institute. When that parent organization combined with the Society of Wireless Telegraph Engineers to form The Institute of Radio Engineers in May, 1912, he continued to attend the meetings and was elected to associate membership November 4th, 1912. He was elected to the grade of Member February 8, 1926. In 1915 he was elected Treasurer and continued in that capacity, except for the year 1917, until his death.

Mr. Hubley, throughout his long and helpful association with The Institute of Radio Engineers, gave unsparingly of time and effort in the interest of the Institute. His unremitting attendance at meetings of the Board of Direction of the Institute rendered available his wise counsel and careful analysis of the problems of Institute management, to which he gave his undivided and thoughtful attention.

The Board of Direction as a group has lost a capable and conscientious associate, and as individuals they have lost a close friend.

Mr. Hubley was born May 9th, 1880. He died at 4 A.M. September 19th, 1927.

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ON THE VALUES AND THE EFFECTS OF STRAY CAPACITIES IN RESISTANCE-COUPLED AMPLIFIERS*

By

Manfred von Ardenne and Wolfgang Stoff

(Berlin)

It is a well-known fact that among the stray capacities in a vacuum-tube amplifier the capacity between grid and plate of the tube is of primary importance, especially since this capacity is materially increased by the amplification of the tube. In Germany this effect first was explained by H. Barkhausen.† The corresponding equations, however, have very often not been properly applied. For example, the effective direct grid-plate capacity is given as $(1+A_v)$ times larger than the statically measured C_{gp}. A_v is the actual voltage amplification of the tube arrangement in question. In this case the fact is overlooked that the phase-difference existent between ΔE_g and ΔE_p is generally not 180 deg., but some other considerably different angle. This fact, of course, is immaterial in respect to mere amplification, but it is of first importance in regard to the charge and actual amount of grid-plate capacity increase. The fact that the actual increase of C_{gp} is, according to the phase angles, smaller than $(1+A_v)$ times, is naturally of the greatest influence on the judgment of amplifier-dimensions such as are given by the writer since the beginning of 1925, namely, resistance-coupled audio-amplifiers with tube-plate resistances and outer plate circuit resistances of the order of megohms. In the following, the general theory of stray capacities will first be developed with due consideration to the phase conditions mentioned above, and then the values and effects of stray capacities will be discussed for a three-stage amplifier of practical dimensions.

Fig. 1 shows the simplified circuit and Fig. 2 the corresponding symbolic diagram employed to explain the calculation of the effective stray capacities in a single-stage amplifier with optional plate load. Here the tube is considered to be an alternating current generator with the e.m.f. $=\mu \cdot \Delta E_{g0}$ and the internal resistance

* Original Manuscript received by the Institute, July 6, 1927. Revised Manuscript received by the Institute, August 12, 1927.

† See also Manfred von Ardenne and Wolfgang Stoff: "Die Berechnung der Scheinkapazitäten bei Widerstandsverstärkern," Jahrbuch der drahtlosen Telegraphie und Telephonie, Vol. 30, No. 13.

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r_p, according to a well-known definition. Parallel to the plate-circuit impedance load Z_b are the plate-filament capacity C_{pf} and the grid-plate capacity C_{gp}, which is connected in series with the

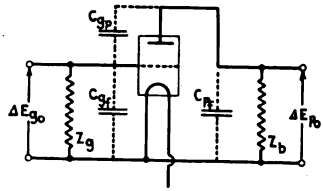

Figure 1

e.m.f. $= \mu \cdot \Delta E_{go}$. The alternating grid-potential ΔE_{go} should be considered as given; ΔE_{po} is the potential across $Z_{b_{tot}}$, which is the total plate-circuit impedance formed by Z_b and the parallel stray capacities. The potential ratio $\frac{\Delta E_{po}}{\Delta E_{go}}$ is given by the following well-known equation:

$$\frac{\Delta E_{po}}{\Delta E_{go}} = -\mu \cdot \frac{1}{1+\frac{r_p}{Z_{b_{tot}}}} \tag{1}$$

Here $\frac{1}{Z_{b_{tot}}}$ is equal to: $\frac{1}{Z_b} + j\omega C_{pf} + j\omega C_{gp} \cdot \left(1 - \frac{\Delta E_{go}}{\Delta E_{po}}\right)$ (2)

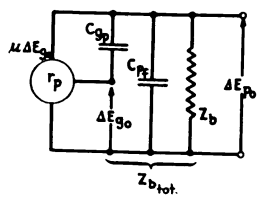

Figure 2

After a mathematical operation with these two equations the following results:

$$\frac{\Delta E_{po}}{\Delta E_{go}} = -\mu \cdot \frac{1 - j\omega C_{gp} \cdot \frac{r_p}{\mu}}{1 + r_p\left(\frac{1}{Z_b} + j\omega C_{pf} + j\omega C_{gp}\right)}$$

Just as in the plate circuit the stray capacities must also be considered in the grid-circuit. Fig. 3 shows the corresponding symbolic diagram for the grid-circuit. Parallel to the grid circuit

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impedance Z_g of optional type are the grid-filament capacity C_{gf} and the grid-plate capacity C_{gp}, which is in series with the e.m.f. $=\Delta E_{po}$. The total grid-circuit impedance $Z_{g_{tot}}$ may then be calculated according to the following equation:

$$\frac{1}{Z_{g_{tot}}}=\frac{1}{Z_g}+j\omega C_{gf}+j\omega C_{gp}\left(1-\frac{\Delta E_{po}}{\Delta E_{go}}\right) \tag{3}$$

In order to make use of equation (3) properly the phase-angle ϕ between ΔI_{po} and ΔE_{po} and ψ between ΔI_{po} and ΔE_{go} have to be introduced. The potential ratio $\frac{\Delta E_{po}}{\Delta E_{go}}$ may then be divided into the real component $\frac{\Delta E_{po}}{\Delta E_{go}}\cos(\phi-\psi)$ and the imaginary one $j\frac{\Delta E_{po}}{\Delta E_{go}}\sin(\phi-\psi)$:

Equation 3 then becomes the following:

$$\frac{1}{Z_{g_{tot}}}=\frac{1}{Z_g}-j\omega C_{gf}\cdot j\frac{\Delta E_{po}}{\Delta E_{go}}\sin(\phi-\psi)+j\omega C_{gp}\left(1-\frac{\Delta E_{po}}{\Delta E_{go}}\cos(\phi-\psi)\right) \tag{3a}$$

In this formula the total effective stray capacity corresponds to:

$$C_{eff}=C_{gf}+C_{gp}\left(1-\frac{\Delta E_{po}}{\Delta E_{go}}\cos(\phi-\psi)\right) \tag{4}$$

For the effective stray capacity only the real component in the ratio $\frac{\Delta E_{po}}{\Delta E_{go}}$ is of influence.

According to the type of plate circuit load three distinct cases can be distinguished for the reaction on the grid-circuit:

First, Z_b is a pure ohmic resistance,

Secondly, the inductive component in Z_b is dominant,

Thirdly, Z_b is mainly capacitive.

For a pure ohmic load on the plate circuit the maximum value of effective stray capacity results, since the first-mentioned formula is correct in this special case:

$$C_{eff_{tot}}=C_{gf}+C_{gp}(1+A_v) \tag{4a}$$

The condition of a pure ohmic resistance load, however, is only met in practice for the lower frequencies (below about 500 cycles), especially in resistance-coupled amplifiers. Furthermore, it should be noted that even if a pure ohmic resistance is connected in the plate circuit, for radio frequencies the stray capacities already

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existent according to equation (2) form a considerable capacitive load.

Generally the effective stray capacities should always be calculated according to equation (4); it then is of no consequence

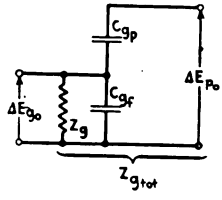

Figure 3

whether the plate circuit has a capacitive or inductive load. It is, however, of importance that, as formula (3a) shows, the imaginary component of the ratio $\frac{\Delta E_{po}}{\Delta E_{go}}$ gives rise to a real component in the grid-circuit. This real component is negative in the case of an inductive plate circuit load, and can therefore produce self-oscillation of the amplifier. For a capacitive plate circuit load this component is positive and therefore damps the grid-circuit somewhat.

In dimensioning audio frequency amplifiers with resistance-capacity coupling, it is of primary importance to determine the frequency response toward the higher audio frequencies, where the capacitive load on the plate circuit is no longer negligible[1]. Correct results are then only obtained if the effective values of the stray capacities are calculated according to formula (4). Besides, the real component appearing in the grid-circuit must be taken into consideration as well, as was mentioned above. This real component has, since all the phase-angles in the formulas given are defined with respect to the direction of the grid-potential as a normal, no phase-angle difference from a pure ohmic resistance in the grid-circuit.

This real component thus effects a reduction of the total grid-circuit impedance $Z_{g\,tot}$ for the higher frequencies; in practice, it therefore is necessary to reckon with slightly smaller grid-leak values in order to take account of this effect. If both effects are taken into consideration: First, the reduction of the effective value

[1] A sufficiently uniform response of the amplifier to impulses of frequencies near the lower limit of audibility may easily be obtained by suitably dimensioning the grid-coupling capacities.

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of the stray capacities because of the phase-difference, and, secondly, the reduction of the total grid-circuit impedance due to the real resistance component introduced into the grid circuit at higher frequencies, then the actual frequency response of the amplifier at the higher audio-frequencies turns out to be considerably more uniform than should be expected from formula (4a).

The practical appliance of the theoretical considerations given above can best be shown by means of an example. Fig. (4) shows the simplified circuit diagram of a three-stage resistance-capacity-coupled amplifier, as will be considered given for the calculations. The dimensions of the components correspond to measurements taken by the writer on an amplifier of this type.

$$R_{b1} = R_{b2} = 3 \text{ megohms}$$
$$R_{b3} = r_{p3}$$
$$R_{g2} = R_{g3} = 8.5 \text{ megohms}$$
$$\mu_1 = \mu_2 = 31$$
$$\mu_3 = 5$$
$$r_{p1} = r_{p2} = 0.7 \text{ megohms.}$$

In a normal circuit arrangement the direct inter-electrode tube capacities for German tubes were measured and found to be from

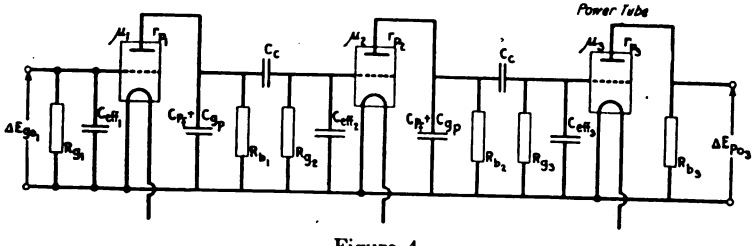

Figure 4

1 to 2 micro-microfarads. In the following a mean value of 1.44 micro-microfarads will be assumed for C_{gp}, C_{gf}, and C_{pf}.

For the simplified assumption of a pure ohmic plate circuit resistance of the power-tube $R_{b3} = r_{p3}$ the following equation exists:

$$\frac{\Delta E_{po3}}{\Delta E_{go3}} = -\frac{\mu}{2} = -2.5$$

The further calculations will be carried out at the frequency of 10,000 cycles which is near the upper limit of audibility.[2] At this frequency the following may be assumed:

$$Z_{b_{tot1}} = Z_{b_{tot2}}; \quad Z_{b_{tot2}} = Z_{g_{tot3}};$$
$$\Delta E_{po1} = \Delta E_{go2}; \quad \Delta E_{po2} = \Delta E_{go3}.$$

[2] At this frequency, the capacitive load of course has the greatest influence.

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According to formula (4a): $C_{eff}=1.44+1.44(1+2.5)=6.5$ micro-microfarads. Accordingly, the total effective stray capacities are found to be $C_{eff_{tot}}=C_{pf}+C_{gp}+C_{eff}=9.4$ micro-microfarads. If the factor $j\omega C_{gp}\frac{r_r}{\mu}$ is disregarded in comparison to (1), then formula (1a) is changed into:

$$\frac{\Delta E_{po}}{\Delta E_{go}}=-\mu\frac{1}{1+r_p\left(\frac{1}{R_b}+\frac{1}{R_g}+j\omega C_{eff_{tot}}\right)}$$

After the correct values for the different constants are introduced:

$$\frac{\Delta E_{po2}}{\Delta E_{go2}}=-31\frac{1}{1+0.32+j\,0.41.}$$

$$\frac{\Delta E_{po2}}{\Delta E_{go2}}\cos(\phi-\psi)=-21.6;\quad \frac{\Delta E_{po2}}{\Delta E_{go2}}\sin(\phi-\psi)=6.8;\quad A_{v2}=22.6.$$

The static amplification in this stage is $A_v=23.8$; the dynamic amplification therefore is practically equal to the static even at the high frequency of 10,000 cycles. The frequency response in this second stage therefore may be assumed to be uniform. After examining the curvature of the dynamic working characteristic it will be found that here too the amplitude-distortion is negligible, which is of primary importance in this stage because the amplitude is large.

For the first stage the following values must be assumed:

$$C_{eff_{tot}}=3\times1.44+1.44\times(1+22.6)=38.5 \text{ micro-microfarads}$$

$$\frac{1}{R_{g_{tot1}}}=\left(\frac{1}{8.5}+0.62\right)10^{-6}=0.715\times10^{-6}$$

Just as above, the following values result:

$$\frac{\Delta E_{po1}}{\Delta E_{go1}}\cos(\phi-\psi)=-9.6;\quad \frac{\Delta E_{po1}}{\Delta E_{go1}}\sin(\phi-\psi)=8.9;\quad A_{v1}=13.1.$$

In spite of the fact that the phase-difference $(\phi-\psi)$ is rather large in this stage (about 140 deg.), the only pure resistance-capacity coupled stage, the dynamic amplification for the frequency of 10,000 cycles does not drop even to one-half of the static amplification. The same of course applies to the three-stage amplifier as a whole:

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$$A_{vst_{tot}} = 1420$$

$$A_{vdyn_{10000}} = 730 = 0.51 A_{vst_{tot}}.$$

These voltage amplification factors given here are those determined by measurement of the input alternating grid potential and of the output alternating potential across the output resistance. The amplification factors calculated here practically exactly comply with measurements taken on a three-stage amplifier of the same dimensions. According to these measurements the dynamic amplification at the frequency of 10,000 cycles falls off to 0.516 times the maximum amplification. Therefore, the effective stray capacity values calculated here may be assumed to be practically equal to those actually present.

As the above calculations have shown, the actual amplification factors of such resistance-capacity-coupled amplifiers and all other amplifiers with appreciable reaction of the plate-circuit load on the preceding grid-circuit can only then be correctly calculated, *if the different stages are calculated proceeding backward, that is, beginning with the plate circuit of the last stage.* Wrong results are obtained, if founded on some assumptions one stage only is examined, and the result gained simply multiplied by the number of stages involved, as in the case of the frequency-response calculation of a three-stage amplifier with resistance coupling. As has been shown by the calculations carried out above, with resistance-capacity-coupled audio-frequency amplifiers the frequency response of the *first stage only* is of material importance. Three-stage resistance-coupled amplifiers of these dimensions with such high amplification factors, however, are seldom used as pure audio-frequency amplifiers. As a rule the grid of the first stage is directly connected to a tuned radio-frequency circuit, which is then coupled either to the antenna or to the output of a radio-frequency amplifier. By the influence of the considerable capacitive load on the plate-circuit of the first tube, a very sensitive "anode-bend" detection effect for radio frequency only is then introduced.[3] Two-stage amplifiers of this description, containing one voltage-amplification stage followed by a power-tube, have a practically complete uniform frequency-response, as is apparent from the calculations given above.

[3] Compare the first writer's publication "Über Anodengleichrichtung," Jahrbuch der drahtlosen Telegraphie und Telephonie, Zeitschrift für Hochfrequenztechnik, Berlin. Vol. 29. No. 3.

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MOUNTING QUARTZ OSCILLATOR CRYSTALS*

By

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Quartz oscillators are becoming more widely used to control the frequencies of radio stations. Under fairly constant temperature conditions, and with good regulation of filament current and plate voltage on the controlled tube, a quartz master oscillator may readily be kept to within 50 parts in a million of an assigned frequency.

There are several methods of cutting a piezo-electric plate from a quartz crystal. Fig. 1 shows a right prism of quartz with the optic axis perpendicular to the bases. The usual orientation[1] for a quartz piezo-electric plate is that of method 2, giving a thickness wavelength of 105 to 115 meters per millimeter. The crystals used

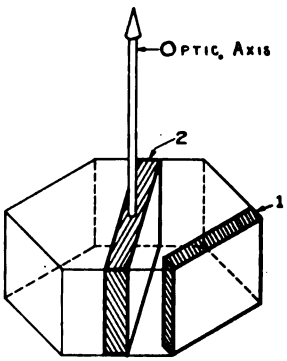

Figure 1—Methods of cutting piezo-electric plates.

for all the tests to be described in this article are cut according to method 1, having a thickness wavelength of 140 to 150 meters per millimeter. For a given wavelength, the crystal of method 1 is two-thirds as thick as a crystal of method 2. For some wavelengths, the thinner crystal of method 1 has definite advantages, due to the stronger electric field which is applied.

* Original manuscript received by the Institute, June 21, 1927.

[1] A. Hund: Uses and Possibilities of Piezo-electric Oscillators. Proc. I.R.E. Aug. 1926, p. 448. A. Crossley: Piezo-electric Crystal-controlled Transmitters. Proc. I.R.E. Jan. 1927, p. 13.

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Of the several factors which govern the action of a piezo-electric crystal when controlling a radio-frequency circuit, the mounting of the crystal is quite important. It will be assumed that the circuit is properly designed, that a good piece of piezo-electric material is available, and that the only variable to be considered is the mounting of the crystal. The main feature of a crystal-controlled master oscillator is its ability to regulate a radio frequency very closely. In addition, it is quite desirable to have the crystal-controlled circuit radiate as much power as possible, in order to reduce the number of radio-frequency power amplifiers.

In using crystal-controlled tubes radiating more than a few watts, several problems arose which did not occur when the power was only a fraction of a watt. The circuit of Fig. 2 is a typical crystal-controlled circuit in which the grid bias is supplied through a

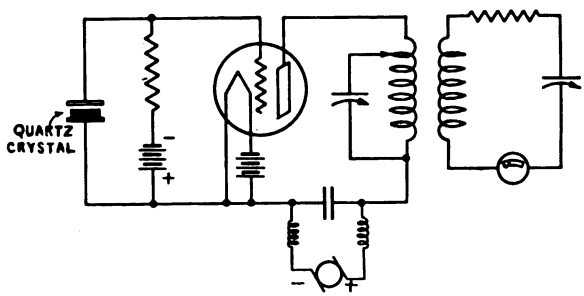

Figure 2—Piezo oscillator with power-absorbing circuit.

grid-leak, this being found to give better control of the radio frequency than when a choke coil was used. In addition to the crystal circuit, an absorbing circuit is used (see Fig. 2 at the right) to measure the amount of power radiated. Using this circuit the various mountings were tried and their effect on the output determined. The frequency characteristics were found by using a crystal-controlled receiving tube circuit coupled very loosely to the power circuit which was being tested. The crystal which controlled the receiving tube circuit was ground to give a frequency several hundred cycles above or below the power crystal circuit, so that the audio beat could be heard by receivers which were inserted in the receiving tube plate circuit.

When the crystal-controlled circuit was arranged so that as much as 10 watts could be absorbed, it was found that the piezo-electric reaction of the crystal was so great as to cause sparking at the electrodes. The sparking caused electrode corrosion which soon interfered with the free action of the crystal, and also spoiled

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the clear-cut character of the radio-frequency output. High power also caused the crystal to heat badly, often resulting in cracking, which of course stopped the oscillations.

In order to compare directly the various types of mountings the frequencies of the crystals were ground to less than 3 percent of one million cycles per second (300 meters wavelength). From a survey of mountings for crystals of this frequency, the possibilities at other frequencies can be considered.

All mountings for piezo-crystals consist of electrodes which are smooth and plane on the side next to the crystal. One of the simplest mountings is that of a flat metal plate on which the crystal rests, and an upper plate resting on top of the crystal, (see Fig. 3). This mounting as shown is one in which the upper plate rests on the crystal, the only pressure being its weight, while others use a spring to press the upper electrode more firmly on the

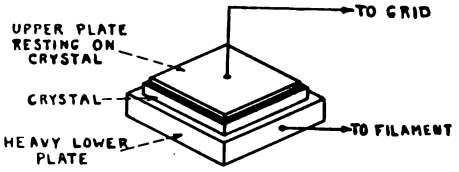

Figure 3

crystal. At high radio frequencies, perhaps above two million cycles per second (less than 150 meters wavelength) the simple type of mounting shown in Fig. 3 is fairly satisfactory.

For frequencies in the vicinity of one million cycles another type of mounting, in which the upper electrode is supported a definite distance away from the crystal, has certain advantages which the simpler mounting does not have. The disadvantages of the simpler holder for frequencies of the order of a million cycles per second are shown when a large amount of power is taken from the crystal-controlled circuit. An ordinary crystal-controlled UX-210 tube (7-1/2 watt rating) can radiate 5 to 7 watts at a million cycles with nearly any type of mounting. When using a UV-211 tube (rated at 50 watts) with 1,000 volts on the plate and absorbing 50 to 80 watts from the plate circuit, crystals were often broken when using the above simple method of upper electrode resting on the crystal. Using the spaced mounting, in an oil bath at constant temperature, which will be described later, tests of several hours were run successfully at outputs of 50 to 80 watts without breaking the crystal.

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Spaced Upper Electrode

When mounting the upper electrode a fixed distance away from the crystal, it is essential to have the electrodes parallel. If the electrodes were not parallel, the electric field would be unevenly distributed, tending to strain the crystal. Various mountings were constructed which provided a space above the crystal, yet which were simple in construction. One type, which proved useful for several sizes of crystals, is shown in Figs. 4 and 5. In this mounting an accurately turned hard rubber ring spaces the upper electrode a small distance away from the crystal, and when a different crystal

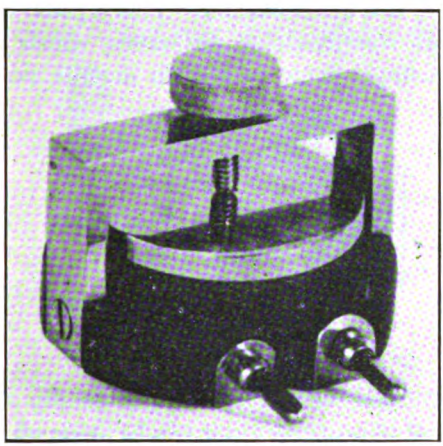

Figure 4

is to be used an appropriate ring is substituted in the holder. A definite spacing which could always be duplicated was afforded by keeping the ring and its crystal in the same envelop.

The crystal vibrations are not damped by the pressure of the upper plate in the spaced type of mounting; this accounts for the increased power which can be obtained. A further advantage of a spaced holder is that the intensity of the electrostatic field across the crystal is reduced, due to what is effectively a series condenser. This reduced field helps prevent breakage of the crystal due to dielectric failure of the quartz.

There is a certain phenomenon which occurs when using a spaced upper electrode which may give trouble if its characteristics are not known. The position of the upper electrode for maximum power and accurate frequency control is somewhat critical. If a

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holder with an adjustable upper electrode is available, an interesting experiment can be performed. If the upper electrode is continuously moved away from the crystal, the radio frequency emitted by the crystal-controlled circuit is raised, but not at a uniform rate. There will be successive points where the frequency will "jump" a few hundred cycles, or occasionally as much as a few thousand cycles per second. If the power is measured at the same time that the frequency is being investigated, it will be found to be a minimum at the same points where the frequency jumps occurred. This test would make it seem that the use of a spaced upper electrode was undesirable, but there is a simple explanation and solution to this phenomenon. The crystal when vibrating strongly sets up compressional air waves, which when reflected back from the upper plate in certain definite positions exert a

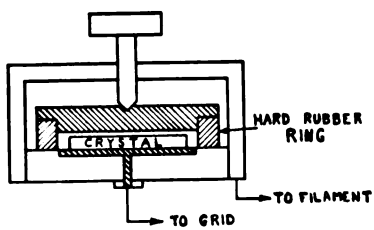

Figure 5

damping effect on the crystal. These are simply standing waves in air. For a first approximation it can be assumed that the speed of this air wave is the same as that of an ordinary sound wave in air. The air wavelength in millimeters would then be equal to the speed of sound in meters per second divided by the frequency in kilocycles per second. That is

$$\lambda_{mm} = \frac{331_{m/sec}}{F_{kc/sec}}$$

It was found that the assumption of the ordinary speed of sound in air did not give the correct distance of air wavelength for minimum power and frequency shift. In fact several experimental curves showed that the speed of the sound wave emanating from a vibrating crystal was nearly twice that of an ordinary sound wave. This makes the wavelength twice as long as when using the usual speed of sound in air, and a more correct formula is

$$\lambda_{mm} = \frac{662_{m/sec}}{F_{kc/sec}}$$

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It should be noted that this equation is simply a further approximation, and that the speed of sound is really a function of the energy with which the crystal vibrates and varies with the amount of power radiated. The reason for the faster speed of propagation is to be found in the very energetic crystal vibration when radiating large amounts of power from the controlled circuit. It is well known that explosive waves travel faster than ordinary sound waves which have much smaller amplitudes.[2] A strongly vibrating quartz oscillator crystal can be regarded as a source of rapidly recurring explosive waves, therefore having a velocity greater than ordinary sound waves.

A simple experiment with a screw top holder serves to locate the correct position of the upper electrode for any piezo-electric

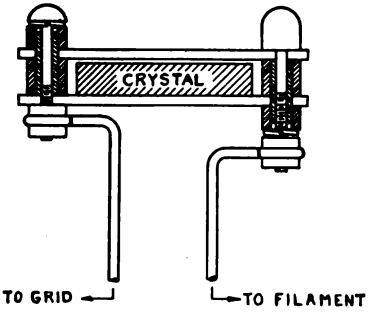

Figure 6—Piezo crystal mounting for low power.

crystal. An absorbing circuit is coupled to the crystal-controlled circuit, and as the upper electrode is raised, the first point of minimum power is noted. The best operating position of the upper electrode is to be found between the upper surface of the crystal and the first position of minimum power. For crystals vibrating at a million cycles per second the spacing of the upper electrode above the crystal can be from two-tenths to four-tenths of a millimeter, the smaller value being preferable.

In mounting electrode plates a fixed distance apart, a reliable spacer should be used. Also, due to the fact that the assembly for permanent use has to be mounted inside a glass bulb, it is seen that any material which contains gases would not be satisfactory. This excludes the use of molded materials such as micarta and bakelite. The heat which is applied to outgas the crystal electrodes is sufficient to rule out other materials which become plastic at

[2] The Detonation Wave from Solid Explosives: W. C. Holmes, *Jour. Fr. Inst.* Apr. 1927, p. 549.

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fairly low temperatures, for instance, hard rubber. In the assembly which is shown in Figs. 6, 7, and 8, the spacers are made of fused quartz tubing. This material has a very low coefficient of expansion, contains no more gases than the glass of the bulb itself, and does not become plastic at ordinary temperatures. Sections of glass tubing can also be used for spacers, although their coefficient of expansion is much higher than for fused quartz. The feature of low expansion coefficient is not of such importance where the

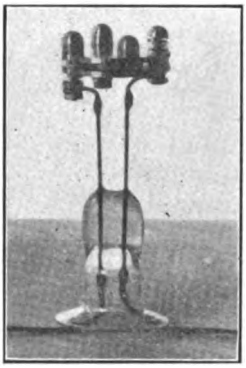

Figure 7

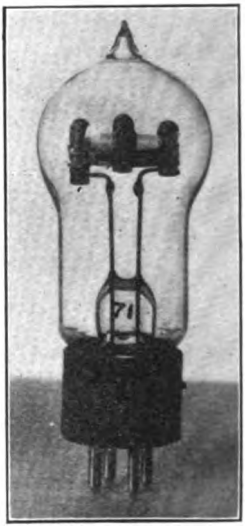

Figure 8

crystal mounting is immersed in a temperature-controlled bath. The oil bath will be described later in detail.

Electrode Material

Having determined the best spacing of the upper electrode above the crystal, the next problem was to find the best material of which to make the electrode plates. When large amounts of power were absorbed from the crystal-controlled circuit, the periodic reaction of the crystal produced such large potential differences between the upper face of the crystal and the lower surface of the upper electrode, that sparks passed between them, as mentioned at the beginning of this article. These sparks are detrimental in two ways. In the first place they are strongly oxidizing sparks which rapidly corrode the electrode plates. Using

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brass electrodes, it was found necessary to clean off the corrosion nearly every day when a crystal was in continuous operation. The other objectionable feature is that air is a fairly good insulator; the spark does not pass until the potential difference has become quite large, and as the spark leaves it creates a disturbance in the form of a transient vibration of the crystal, which is added to the emitted radio frequency. When a crystal plate is sparking, the rough pitch of the sound can be heard by listening near the crystal holder, and this pitch is the same as that which is added to the radio frequency, making it unsuitable for master oscillator control of radio stations. To reduce the effect of oxidizing the electrodes, metals which were somewhat resistant were adopted, such as

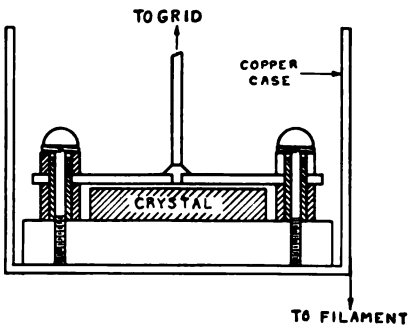

Figure 9—Piezo crystal mounting for high power.

monel metal, pure nickel, and stainless iron. Of these three, pure nickel is perhaps the easiest to machine and makes a very good crystal electrode.

There are two methods of procedure which can be used to prevent sparking. The original procedure was to reduce the plate voltage of the crystal-controlled tube. The output power was thereby reduced, and in turn the voltage feedback through the tube was diminished, which eliminated the sparking at the crystal. The disadvantage was obvious; the number of radio amplifier stages had to be increased to compensate for the decreased power in the crystal-controlled tube circuit.

Glow Discharge at Crystal

The second method is the one which is now being used successfully for control crystals at the radio station KDKA. This method substitutes a silent glow discharge for the rough spark. The glow discharge takes place at much lower potential differences than the spark, and accomplishes the desired result. A glow dis-

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charge is best accomplished by using some gas which ionizes readily, which means that the crystal holder must be sealed, as in Fig. 8. The glow discharge is preferably carried by an inert gas, so that it will not attack the electrode plates when in its ionized state. Neon is an inert gas which permits glow discharges at very low potential differences, and a pressure of a few millimeters of neon has proved satisfactory for quartz oscillator plates.

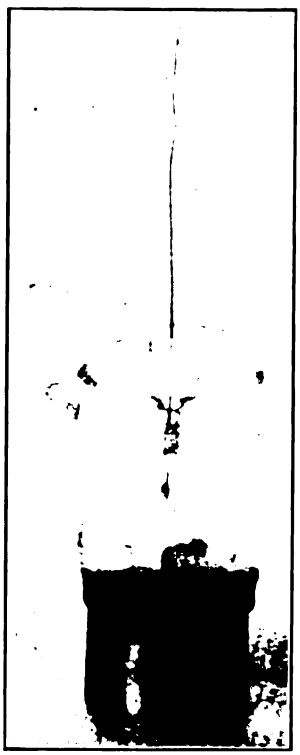

Figure 10

UX-210 tubes with 400 volts on the plate can radiate 5 to 7 watts of clear-cut radio-frequency output when using a crystal plate mounted in a neon-filled tube. Under the same conditions of clear radio output, the same crystal in air would only be able to supply one or two watts and the tube would have to be operated at a reduced plate voltage of 200 or 300 volts.

When piezo crystals mounted in neon are controlling power circuits, there can be seen a pink glow surrounding the crystal. If the plate circuit is tuned to a frequency much higher than that

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of the crystal, a brilliant glow will fill the entire crystal tube. As the frequency of the plate circuit is lowered, the glow becomes more concentrated, until when the plate circuit is tuned to its best point (slightly higher than the crystal frequency) the crystal itself seems to be permeated with the pink glow, but there is very little glow in the rest of the tube.

The effect of varying temperature is a very important one for quartz plates. Depending on the method of cutting the quartz plate, the frequency-temperature coefficient is from 25 to 50 parts in a million per centigrade degree. For quartz plates oscillating in the thickness direction the latter figure should be used. For a constant radio frequency the temperature of the quartz plate should be kept as uniform as possible. Quartz crystals are not symmetrical; it is due to this fact that they are piezo-electric. This asymmetry is a disadvantage in that the crystal when heated expands unevenly in its thickness and length directions. It expands very nearly twice as fast in the thickness direction as it does in one of its longer dimensions. Any use of a crystal to control more than 10 watts causes so much dielectric displacement current to flow through the crystal that it heats up. If the crystal is operated continuously without cooling the heat finally causes such large stresses that the crystal cracks. Sometimes only a corner cracks and again the crack may occur through the middle of the crystal. If only the corner has been cracked, the crystal can be made to oscillate again by carefully grinding away all the portion that shows any cracks. If a large crystal cracks in the middle, it may be cut into smaller plates which are all right for radio oscillators. In other words, the failure seems to be a local one and the rest of the crystal is unaffected.

Cooling Crystals for Power Work

A means of cooling the crystal is necessary when continuous operation at high powers is desired. The holder which is used to do this is shown in Figs. 9 and 10. This mounting retains the features of the glass bulb type, in that the upper electrode is spaced above the crystal, and an inert gas at low pressure is used for a glow discharge. By using direct metallic conduction from the lower electrode to the copper case, which is sealed to a glass superstructure, the cooling is easily effected by immersion in a tank of oil kept at the desired temperature. Fig. 11 shows the tank with the crystal unit immersed, and Fig. 12 gives the connections for the thermostat T, the heating element R, also the lamp L and condenser C which serve to protect the thermostat contacts on

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breaking the circuit. For a 3-gallon oil tank, R was 85 ohms, and L was a 100-watt tungsten lamp. The working temperature of the crystal unit is chosen to be above that of the highest summer room temperature. The temperature chosen for a series of tests was 58 deg. C. A good bimetallic thermostat of the helical type

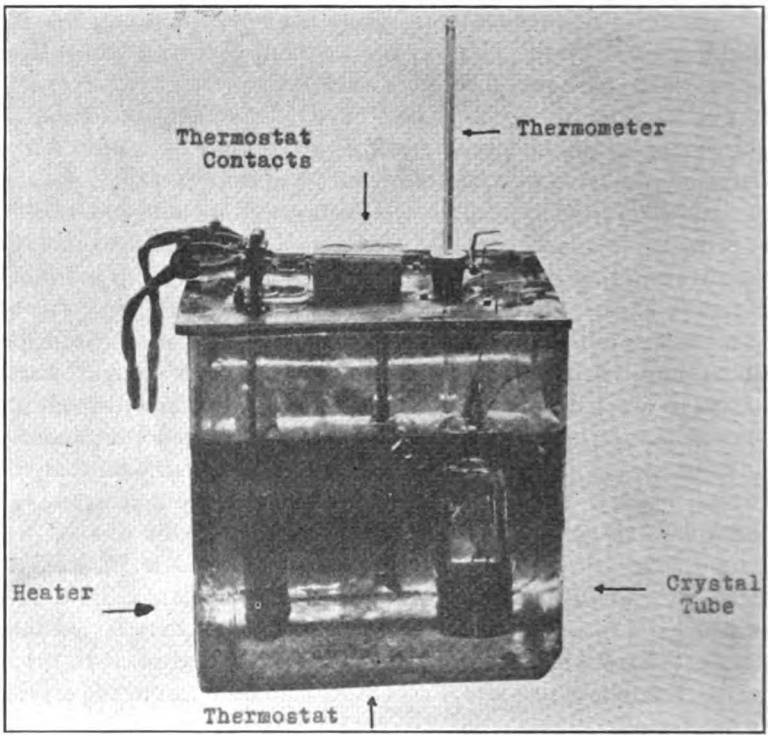

Figure 11

was found to regulate the temperature of this 3-gallon oil bath within two-tenths of a centigrade degree. This variation should keep a piezo-electric quartz crystal constant to better than 10 parts in a million. Frequency differences are difficult to measure to better than 20 cycles when using a million cycles per second, but it can be said definitely that the control is better than 20 parts in a million.

It should be noted that both the heater and the crystal are placed below the center of the oil bath. This is to insure adequate cooling of the crystal and to provide good circulation in the case of the heater. If the crystal heats up above the oil temperature,

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the thermostat cuts off the major part of the current through the heater, and the coolness of the room serves to bring the oil temperature down. If the crystal does not heat up very much, the regular heater is turned on and off by the thermostat, in keeping the oil bath at the desired temperature.

Using a crystal unit in the oil bath mentioned, and feeding into the grid a UV-211 tube using the circuit of Fig. 2, and applying 1000 volts on the plate, a test of 8 hours was run while absorbing 50 watts in the auxiliary circuit. Longer tests would be illuminating and are in progress, but it is felt that if there were a flaw in this manner of operating a crystal, that it would show up

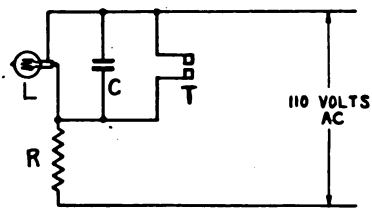

Figure 12—Thermostat and heater connections.

in a severe test of the kind just described. The temperature-controlled oil bath can also be used for crystals mounted as in Fig. 8 which are not used primarily for maximum power output.

For crystals of natural thickness frequencies of about one million cycles per second, a spaced electrode holder, mounted in an inert gas, and properly cooled, is recommended for power work. For crystals of higher frequencies, there may be some difficulty in making the spacing of the upper electrode as small as is necessary to avoid the resonant air wave. In this case, it might prove advisable to use a harmonic of a lower frequency having a large power output. For crystals of lower frequencies than a million per second there should be no trouble in using exactly similar mountings to those of Figs. 6 and 9, with a small distance between the crystal and upper electrode.

Summary

Various types of mountings are described for quartz oscillator crystals having frequencies in the vicinity of one million cycles per second. For low power work a spaced upper electrode is recommended, mounted in a neon-filled tube. For radiating up to 50 watts from the crystal-controlled tube, a similar mounting placed in a copper-based tube is advised. This latter tube can be cooled by immersion in an oil bath. Accurate temperature control of the oil bath keeps the radio frequency within narrow limits.

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THE SHORT WAVE† LIMIT OF VACUUM TUBE OSCILLATORS*

By

C. R. ENGLUND

(Bell Telephone Laboratories, Inc.)

The following paper contains the results obtained while investigating the practical short wavelength limit of the vacuum tube oscillator. While considerable work at short wavelengths has been published by various investigators, no conclusive data determining the practical short wave limit have come to the author's notice.

Short waves may be obtained from vacuum tubes in at least two quite distinct manners, the first being the orthodox oscillator

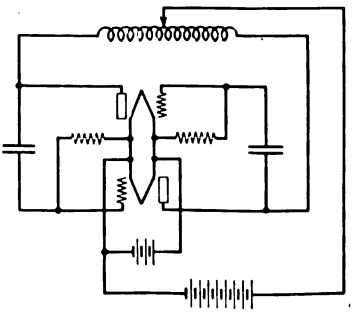

Figure 1

operation where a coupling between grid and plate circuits by regeneration builds up resonance oscillations of frequencies determined by these circuit constants, and the second (the Barkhausen type) being an oscillation within the tube itself in the electron atmosphere between grid and plate. The frequencies of the second type of oscillation are apparently determined by the time of transit of an electron between the tube electrodes and are thus dependent upon the electrode voltages. Little power has so far been obtained in the second manner. The work here reported refers to the first type or normal oscillator operation.

It was found that shorter waves could be obtained by means of a double tube symmetrical oscillator, than could be got by a single oscillator, due to the fact that the total length of conductor carry-

* Original manuscript received by the Institute, July 24, 1927.

† In the present state of the short wave art, it is a distinct inconvenience to use "frequency" rather than "wavelength," since all the measurements are made in terms of lengths.

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ing oscillating currents could be reduced, in the double tube connection, below what could be reached by means of a single tube. For wavelengths above five meters the tubes available in commercial use were found adequate, and except for the limitations imposed by the necessity for low resistance condensers and inductances of very small values, the oscillator technique differed little from that at considerably longer waves. Small inductances of low resistance at these wavelengths require considerable copper, but commercial tubing has been found adequate.[1] This paper therefore concerns only the range below five meters.

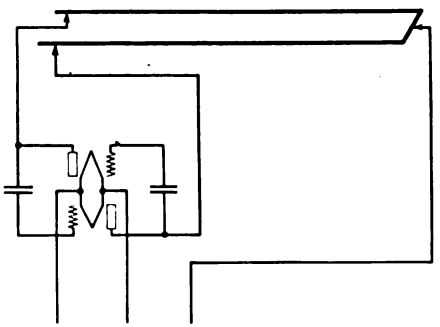

Figure 2

In a cursory consideration of experimental procedure it might seem that the connection of a vacuum tube to a "Lecher System" of parallel conductors offers greater possibilities of reaching shorter waves than does the use of lumped circuits. This has not been borne out by our experience. The first type of circuit tried was a symmetrical connection of standard *215-A* tubes (*N* tubes) Fig. 1, connected to a quarter wave Lecher system as shown in Fig. 2. Results showed that the waves actually obtained were much longer than four times the length of the Lecher system. The following simple theory, while not exact, is adequate to explain the results.

The impedance of a Lecher wire system of length l and shorted at the far end, is given by the expression

$$S_o = \sqrt{\frac{Z}{D}} \tanh \sqrt{DZ}\, l$$

where $Z = R + i\omega L$ is the complex series impedance per unit length and $D = A + i\omega C$ the complex shunt admittance per unit length

[1] The ordinary 1/8 in. (0.318 cm.) commercial copper tubing was chiefly used here.

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of the system. Since the tube reactances are capacities we may say that oscillations will occur when the Lecher wire impedance is an inductance and these oscillations will be approximately of the frequency given by $\omega^2 L'K = 1$ where L' is the effective inductance of the Lecher wires and K the capacity of the tube electrodes in shunt on the Lecher wires. Or

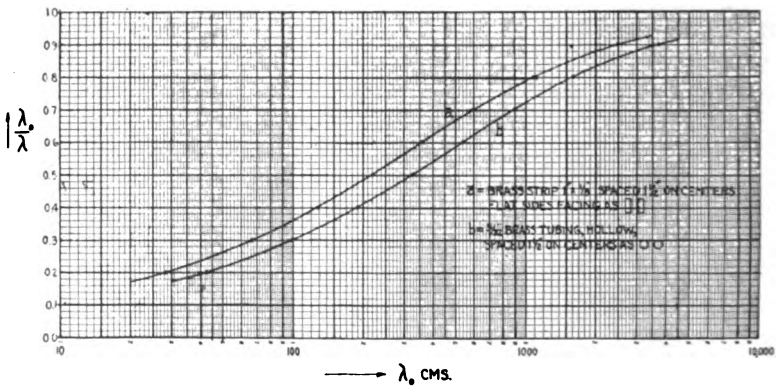

Figure 3

$$\text{imag. part of } \left[\sqrt{\frac{R+i\omega L}{A+i\omega C}}\tanh\sqrt{(R+i\omega L)(A+i\omega C)}\,l+\frac{1}{i\omega K}\right]=0$$

is the resonance condition. Neglecting R and A, since approximate values will suffice, we have the transcendental equation

$\sqrt{\frac{L}{C}}\tan\omega\sqrt{LC}\,l=\frac{1}{\omega K}$ or if λ be the true wavelength obtained and $X=\frac{\pi}{2}\times\frac{\lambda_0}{\lambda}$ $\lambda_0=4l$ we have $X\tan X=\frac{C\lambda_0}{4K}$ which may be readily solved graphically. In order that $\lambda=\lambda_0$ we must have $\frac{Cl}{K}=\infty$ or only at relatively long wavelengths does the ratio $\frac{\lambda_0}{\lambda}$ (for given tubes) approach unity. If $\frac{\lambda_0}{\lambda}$ be determined experimentally for a given wavelength it can be calculated for other wavelengths and Fig. 3 gives two curves thus obtained for two Lecher systems, markedly different in shunt capacity. No essential distinction can

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be drawn between such Lecher systems and a lumped inductance at the shorter wavelengths; it is only at longer wavelengths that space wave resonance can occur. Naturally the addition to the Lecher system of a length of additional wire equal to half a wavelength (or 1/4 wavelength if open circuited at the end) need not affect the wavelength generated and, conversely, by damping out the fundamental wave by means of appropriate shunts at the current antinodes harmonic oscillations should be produced. The additional energy loss is apparently pronounced, however, for

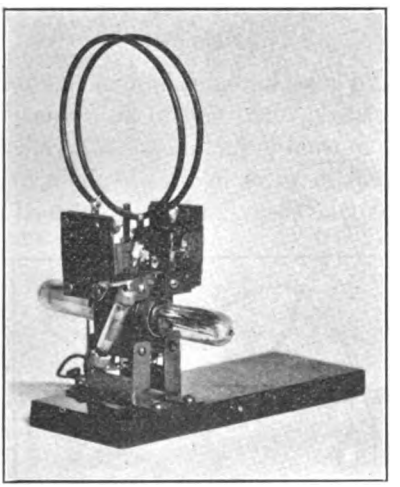

Figure 4—4.2-meter wavelength

while it has been possible to produce such harmonic waves, only the fundamental occurs without shunts. A little consideration will show that no harmonic wavelength is obtainable which is not more easily gotten as the fundamental of a shorter length of line. Accordingly, no further attempts to excite Lecher systems by means of *direct* connected oscillators were made and simple inductances were solely used for the remaining work.

Fig. 4 is a photograph of a balanced *215-A* tube oscillator using the circuit of Fig. 1 and giving a wavelength of 4.2 meters. By reducing the size of the coil this wavelength was considerably reduced. The tube capacity was the only tuning capacity used. Realizing that too much metal was used in this oscillator it was rebuilt with very compact blocking condensers of tinfoil and paraffined paper and a wavelength of 1.9 meters was reached.

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In casting about for some scheme of connecting the tubes so as to reduce their effective capacity the circuit of Fig. 5 was devised and tested. It operated well, usually without a grid leak and could be pushed down to 1.67 meters with ordinary based

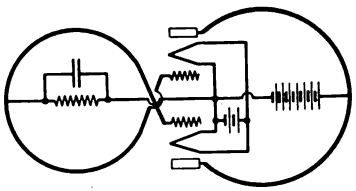

Figure 5

215-A tubes. Fig. 6 is a photograph of the actual oscillator. The two coils (each a single turn) had to be connected geometrically as shown; if their mutual inductance was reversed no oscillations occurred. The lead in wires in *215-A* tubes are about as short as it is possible to make them, and when unbased tubes were used,

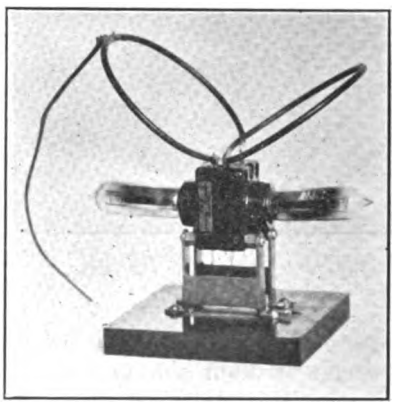

Figure 6—1.67 meters. $E_b = 180$ volts.

as in Fig. 7, the wavelength was reduced only to 1.42 meters.

Since this represented the limit attainable with commercial *215-A* tubes attempts were made to push down in wavelength by increasing plate voltage and filament current. A plate voltage increase is of no use alone if the filament does not have a corresponding emission of electrons, in fact a very active filament is the backbone of a short wave oscillator. The limit is reached when the plate overheats. No marked shortening in obtainable wavelength occurred by pushing the plate current per tube to 17

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milliamperes momentarily (E_b app. 200 volts, normal I_b 2 milliamps.) and had such shortening been obtained it would not have been of practical value.

It therefore appeared that the only prospect of further reduction in wavelength was in the construction of a special tube

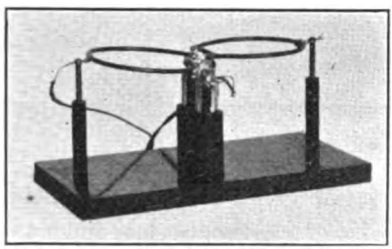

Figure 7—1.42 meters. $E_b = 180$ volts.

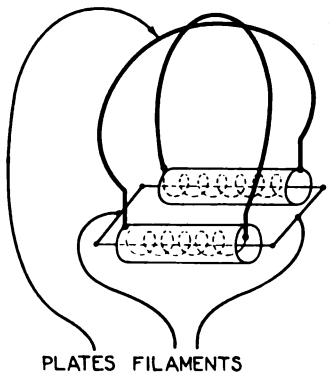

Figure 8

and the assembly adopted was that shown schematically in Fig. 8. Five of these tubes were tested, the last one with doubled leads inside the tube and one-ampere filaments. With this a wavelength of 1.05 meters was finally reached, as a rather unstable

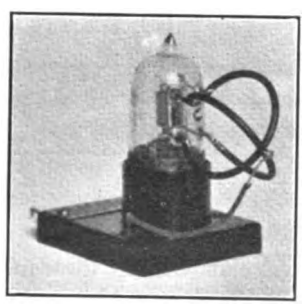

Figure 9—1.05 meters. $E_b = 180$ volts.

oscillation, by pushing the tubes hard. Fig. 9 is from a photograph of the complete oscillator. At two meters the operation of these tubes was quite satisfactory and Fig. 10 shows such an oscillator arranged to vary the frequency by changing the length of the oscillating coils through sliding contactors. The circuit is the same, electrically, as that of Fig. 8.

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As is evident from Fig. 9 the physical dimensions have been reduced to the point where it is necessary, for further progress, to put the resonant circuits inside the tube itself. The inter-electrode capacity of such a tube would have to be kept very small also, all further experiments having indicated that this is more important than low plate impedance. So far no limitations except those of

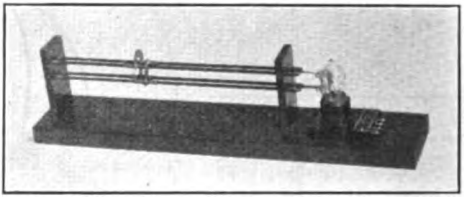

Figure 10—2 meters.

ordinary energy loss have been apparent. But to operate below one meter a tube must have the properties of a low loss condenser with terminal leads of very low inductance. Unnecessary inclusion of glass dielectric and the use of nickel electrodes are inadvisable and the tube geometry must be rigidly worked out. The construction of such a tube is necessarily a double assembly to keep the leads short, and the wavelength is not variable.

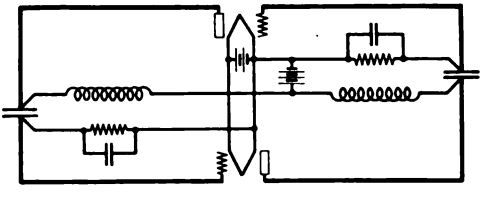

Figure 11

At the shortest wavelength obtained a grid leak stopped the oscillations, at longer wavelengths, say three meters, the oscillations were so vigorous that if a leak was used it had to be a low resistance one or the tube "blocked" at a rate varying from 4 to 10 kilocycles per second. The oscillating current during blocking was very rich in harmonics, and these could be demonstrated as of wavelengths well below one meter. If necessary, therefore, the use of harmonics will enable the production of small amounts of energy at wavelengths below one meter and it is possible that this will be technically useful sometime. It might appear that such harmonics could be amplified after selection but apparently the technique

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which will produce amplifier tubes will give oscillator tubes, therefore the need for amplification brings the problem back to the oscillator itself. Possibly a square law detector pushed hard with an input from a short wave oscillator will make a useful source of shorter waves.

Another double-tube oscillator circuit which seems feasible is given in Fig. 11. It is complicated by requiring blocking condensers and for the shortest waves needs a different geometrical arrangement from that in Fig. 8. It was tried once and not giving any apparent improvement over the circuit of Fig. 8 was not used further.

The one-meter wave not being controllable, it is pertinent to determine what wavelength limit can be reached with reasonable

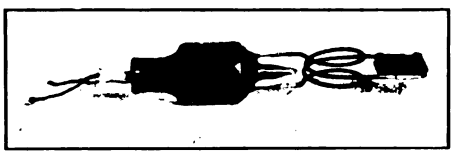

Figure 12—3.1 meters. $E_b = 700$ volts.

power and some frequency control, with the present vacuum tube technique. With the double *215-A* tube a two-meter wave is available with frequency control, with a reservation to be referred to later. While the power thus available is small, it is sufficient for wavelength, capacity, and inductance measurements, a subject which will be deferred for the moment. Some rough tests on commercial tubes, operated at normal voltages are added in the table below.[2] These values are not necessarily the shortest attainable but represent wavelengths which can be obtained with reasonable ease.

Tube	Connection	Meters
230-D (60 mil. fil.)	push pull	2.0
205-D ("E" 5 watt)	single (200 volts E_b)	3.2
221-D (1/4 amp. fil.)	push pull	3.3
211-D ("G" 5 watt)	push pull	3.5

No frequency control was left at these wavelengths. It therefore appeared possible to design a *5-watt* tube which should have some frequency control left at 3.5 meters and a photograph of such a tube is shown in Fig. 12. With the oscillating circuit shown

[2] A recent measurement has given Radio Corporation UX-852, single tube, 1,000 volts 2.6 meters.

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attached, the wavelength is 3.1 meters, and by means of a small variable shunting capacity the wavelength can be controlled within a reasonable range. With such a tube (at 3.66 meters) it has been possible to transmit, without noticeable attenuation, three watts for a distance of 5.2 meters (the room would not allow a longer distance) over a brass pipe transmission line, and radiation experiments, using a simple vacuum tube voltmeter as receiver, have been made up to a distance of 5 1/2 wavelengths. At a four-meter wavelength, reception up to distances of a mile has been possible with a double detection receiver.[3] It seems certain, therefore, that wavelengths down to 3.5 meters are readily available for engineering

Figure 13

Figure 14

Figure 15

purposes. Figs. 13, 14, and 15 show transmitter, receiver, and transmitter reflector antenna systems operating at 3.66 meters. It is a curious fact, possibly worth mentioning, that the human body forms a very fair Hertzian resonator, and at 3.66 meters is so nearly in tune with the radiation that an operator becomes a mobile parasitic antenna seriously interfering with radiation experiments.

The results given above are somewhat at variance with the data published by a number of investigators[4] since only by the use of special tubes was it possible to shorten the wavelength below that reported for (apparently) ordinary tubes by these investigators. Especial care was therefore taken to make certain that the values here reported are fundamental wavelengths and not harmonics of a fundamental but unobserved oscillation. It is not difficult to assure certainty in this matter; it is merely necessary to have a resonant Lecher system which is longer than a full wavelength and which has a low attenuation. Brass tubes of

[3] I am indebted to Mr. H. C. Baumann of this laboratory for the construction of this receiver.

[4] See list of references attached.

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0.8 cm. diameter spaced approximately 7 cm. give such a system and with this it should be difficult to mistake harmonic minima for the practical zeros of the fundamental. Various Lecher frames have been made and compared, both as open end and short circuit end frames. The check between $\lambda/4$ and $\lambda/2$ frames was always found good to 0.2 percent although it would seem that a short circuit end, when in the form of a plate, would be more nearly ideal than an ordinary open end. Figs. 16 and 17 show two short and adjustable $\lambda/4$ frames used, one having a meter in it. The presence of the meter necessitates a small correction. The method of using these frames was found to be very simple. For the shorter

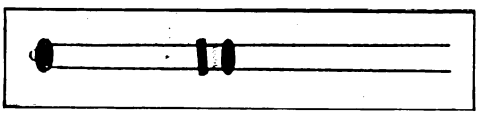

Figure 16—Range 1.2–3.2 meters.

wavelengths it was sufficient to place one end of the wave frame near the oscillator and adjust by observing the tube plate current. At space resonance the energy absorption from the tube circuit was so marked that it could not fail of observation and would stop the oscillation when increased. For longer waves where more power was available a meter in the Lecher frame could be used for the rough setting and for the fine setting a sensitive thermocouple and micro-ammeter connected to a coil of several turns of wire could be used as a loosely coupled exploring unit to indicate maxima and minima on the frame while its length was being varied. Almost equally sensitive is a vacuum tube voltmeter with tube mounted in the handle holding the wire coil. If the experiment is one involving radiation, a Lecher frame in proximity to a sharply tuned receiving antenna and detector gives very distinct and accurate wavelength settings.

The measurement of small inductances and capacities by means of Lecher frames has been mentioned. The theory, when great accuracy is not required, is simple. Thus, suppose a $\frac{\lambda}{4}$ Lecher frame be cut off a distance X from the closed end, the length of open line $\frac{\lambda}{4} - X$ being rejected. Obviously, this frame will be again in resonance to the wavelength λ if an impedance

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equal to the effective impedance of the piece cut off is placed across the open end. The impedance of a short length of Lecher frame with open end is

$$S_\infty = \frac{\sqrt{\frac{Z}{D}}}{\tanh \sqrt{DZ}\left(\frac{\lambda}{4} - X\right)}$$

and if we neglect the resistance and conductance of the frame

$$S_\infty = \frac{\sqrt{\frac{L}{C}}}{i \tan \omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}$$

or as long as $\frac{X}{\lambda} < \frac{1}{4}$ the impedance is capacitative. Hence,

$$\frac{1}{i\omega K} = S_\infty = \frac{\sqrt{\frac{L}{C}}}{i \tan \omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}$$

and

$$K = C\left(\frac{\lambda}{4} - X\right)\frac{\tan \omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}{\omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}$$

is the capacity necessary to reterminate the Lecher frame for space resonance. For small values of $\left(\frac{\lambda}{4} - X\right)$,

$$K = C\left(\frac{\lambda}{4} - X\right)$$

a very convenient expression. When $X > \frac{\lambda}{4}$, or the wave frame has been lengthened instead of shortened, K becomes negative and the impedance which will effectively retune the wave frame to a $\frac{\lambda}{4}$ length is an inductance. This is a particularly neat method of adjusting choke ciols for short wave work, as was experimentally verified.

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A reservation was mentioned in connection with the two-meter oscillation obtained with the apparatus of Fig. 10, with reference to the frequency control. The frequency was readily adjusted to the required value on an ordinary percentage basis, but when a heterodyne reception with two oscillators was attempted it failed. Two meters is the equivalent of 150,000,000 cycles and 1 k.c. is therefore only one part in 150,000 and a setting to this accuracy would be considered difficult under any circumstances. Here, with the tube elements themselves constituting the major part of the tuning capacities, the natural frequency variations were so pronounced that a heterodyne note did not exist; only a hiss could be heard as the oscillator frequencies passed through the

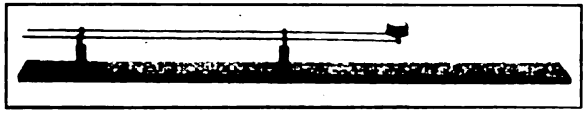

Figure 17—Range 3.4–6.2 meters.

region of synchronism. Later a beat was looked for at 3.6 meters, and this was successful. But the resulting note was extremely variable. Some of this variability was traced to the a-c. excitation of the power tube (full wave rectified alternating current on the plates and straight alternating current on the filaments) where the mechanical tractions set up inside the tube were apparently great enough to wobble the tube elements markedly. But also on battery excitation the frequency variation was considerable, and even when the coupling between the oscillators was sufficient to pull the frequency of one about, as the frequency of the other was varied, a clear note was difficult to obtain. Possibly, with great care, heterodyne reception will be feasible, but it will not be an engineering proposition without something like crystal control.

A list of short wave vacuum tube oscillator references, which it is hoped is complete, is attached.

Summary

An investigation of the short wave limit of vacuum tube oscillators of the normal type, where regeneration builds up oscillations in a resonant circuit, has indicated that the physical limits for ordinary commercial tubes lie between 3.5 and 1.5 meters, depending on the type of tube used. By means of a small special tube a wavelength of 1.05 meters was reached. No frequency control

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was left at this wavelength. With a small power tube not deviating greatly from ordinary vacuum construction it was possible to operate at 3.5 meters with adequate frequency control, and it is concluded that the 3.5–5 meter range is available for technical purposes. Apparatus for measuring wavelength is discussed together with its application to the measurement of capacity and inductance. A bibliography of short wave vacuum tube oscillator work is attached.

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DIRECTIONAL RADIATION WITH HORIZONTAL ANTENNAS*

By

A. MEISSNER

In continuation of the investigation with horizontal antennas and reflectors[1] the attempt was made to concentrate the radiation energy by combining several antennas. If several horizontal

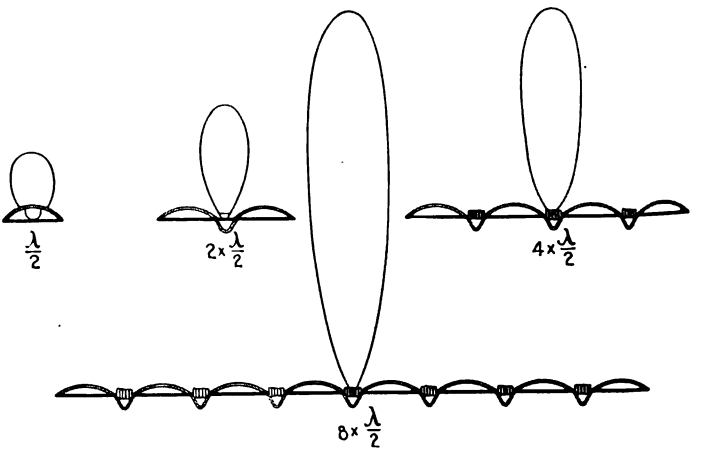

Figure 1—Concentration of radiation in the horizontal plane.

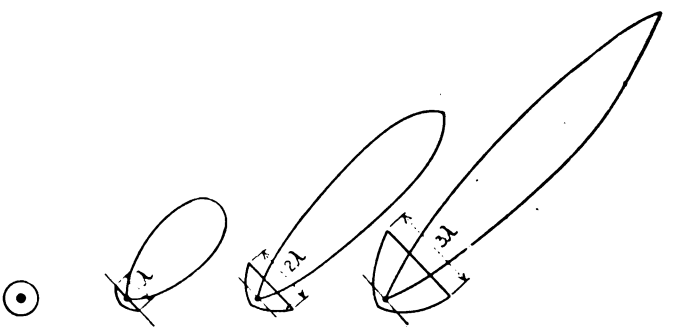

Figure 2—Concentration of radiation in the vertical plane.

antennas are so arranged as to oscillate together in the same phase, radiation patterns are obtained as shown in Fig. 1. The radiation angle as measured in the horizontal plane (angle determined by the

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[1] Jahrbuch für drahtlosen Tel. & Tel., Vol. 28, 1926, page 78.

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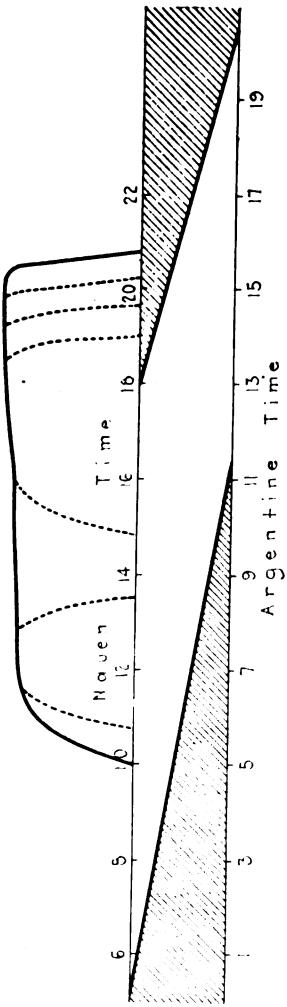

Figure 3—Communication periods with the Argentine on 11 meters.

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half-amplitude of the radiation on both sides of the maximum) diminishes with increasing number of antennas in the ratio of 72 deg.:42 deg.:32 deg. to 14 deg. for 1, 2, 4, and 8 half-wave radiators respectively. The equal phasing of the radiation is obtained by the use of a non-radiating half-wave system[2] (non-radiating loops, coils or combination of coils and condensers) connected between two adjacent antennas. The concentration of the radiation in the vertical plane (Fig. 2) is obtained by an arrangement of a parabolic reflector about the horizontal antennas.

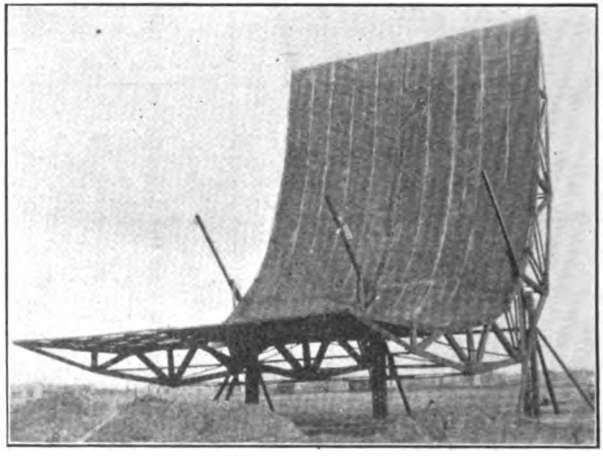

Figure 4

The concentration of the beam is sharper the greater the ratio of the width of the opening of the parabola to the wavelength.[3] The first reflector used was built of copper sheet and other reflector systems were compared with this "standard system." It was found that the metal-sheet surface could be entirely replaced by wire with equally good results; the individual wires of the wire reflector being one-half wavelength long.

With a beam concentrated in this way in both the horizontal and vertical planes, the determination of the optimum angle for the radiation from the transmitter was next attempted. In order to reduce the dimensions and the cost of the entire beam transmitting system as much as possible, the shortest possible wavelength was used at which reception over the distance 10,000 km. (South

[2] Franklin, British Patent.

[3] Tatarinoff, Jahrbuch für drahtlosen Tel. & Tel., Vol. 28, 1926, page 117. It seemed more suitable to deviate somewhat from the parabola form.

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America) could still be expected. According to computations and theories[4] no reception would in general be possible with 11 meters over greater distances, as according to theory the beam is bent away from the earth at such high frequencies. Contrary to theory, the investigation showed that not only did the wavelength of 11 meters work extraordinarily well over great distances, but that this wave when horizontally polarized by reflectors was for the most part even the best for daylight transmission with South America. This showed that the wavelength range for short wave communication could be extended considerably downward. In connection with reception in South America, several interesting

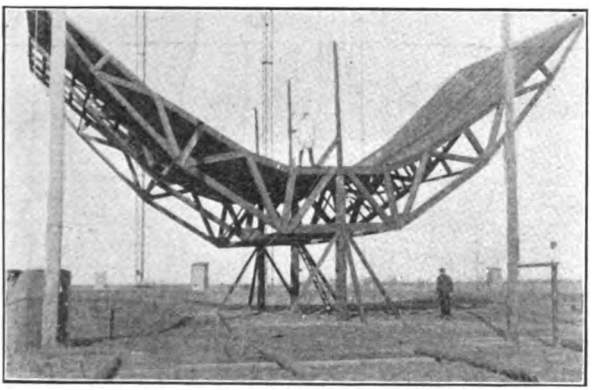

Figure 5

things were noted. The reception on the wavelength of 11 meters was also possible during the period in which a part of the intervening path lay in darkness. Fig. 3 shows the time of reception; underneath, the period of daylight for the month of April (above for Nauen and below for Rio de Janeiro). The reception began usually about 10 o'clock Central European Time. The period from 13 to 15 o'clock is often one of uncertainty. From then until 21 o'clock reception is very good. It was further characteristic of 11-meter transmission that on certain days reception failed completely. The following values of audibility ratios were obtained in comparing transmission from the 11-meter transmitter *AGK* antenna energy less than 1 kw. and the 15-meter transmitter *AGA*, antenna energy 8 kw. (beam radiator with vertical polarization;

[4] A. H. Taylor and E. O. Hulburt, Phys. Rev., Vol. 27, 1926, page 189. Lassen, Jahrbuch für Drahtlosen Tel. & Tel., Vol. 28, 1926, page 109.

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8 parallel wires half a wavelength long and half a wavelength separation):

Date	Time	*AGK* (1 kw.) =11 m.		*AGK* (8 kw.) =15 m.
May 17, 1927	14–19 o'clock	1	:	2–6
	19–21 "	1	:	1
May 18, 1927	14–20:30"	5	:	1
May 19, 1927	14–19 "	6	:	1
	19–20:30"	2	:	1
May 5, 1927	none		normal	

The transmitter *AGK* is, in other words, better than the transmitter *AGA* but at times fails completely. It appeared as if the

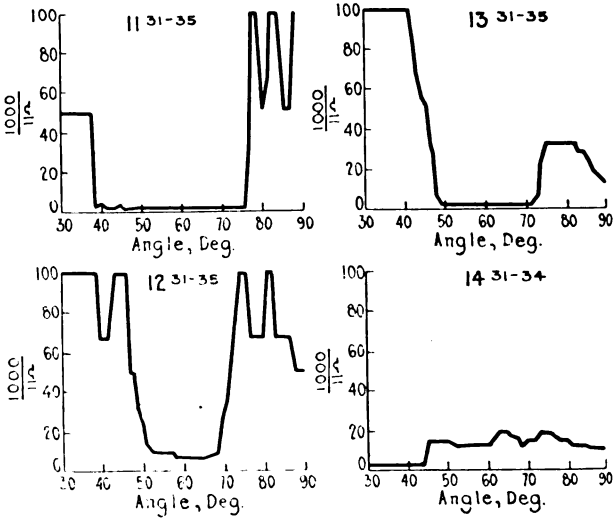

Figure 6—Signal strength for various reflector angles.

latter was somehow connected with a depression approaching the transmitter. Should this be the case, that the influence of the lower atmosphere (up to 20 km.) is of significance, a high reflector angle (80 deg.-second optimum as shown later) may bring the signal back in again on such days that the signal with low reflector angle goes out completely. The investigation so far had only been carried out with a radiation angle of the reflectors of 38 deg.

Reception on 11 meters shows comparatively less fading than on 15 meters. How much of this decrease in fading might be due to using 500-cycle plate supply on the 11-meter transmitter is not known. Fig. 4 shows the beam reflector system which was used. The antenna consists of two half-wave sections connected by a phase reversal coil. To this coil are connected the conductors

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leading from the transmitter which is located 28 meters distant. The reflector is 11.5 meters long and 19 meters wide and rests on three supports 3.9 meters high. At the right in the figure are shown the two windlasses with which the reflector is turned. Two men, one on each windlass, can easily raise the reflector from 30 deg. to 90 deg. in four minutes. Fig. 5 shows the reflector in the 90 deg. position. (The transmitting arrangements were in charge of Mr. J. Pohl, while the reception observations were made by

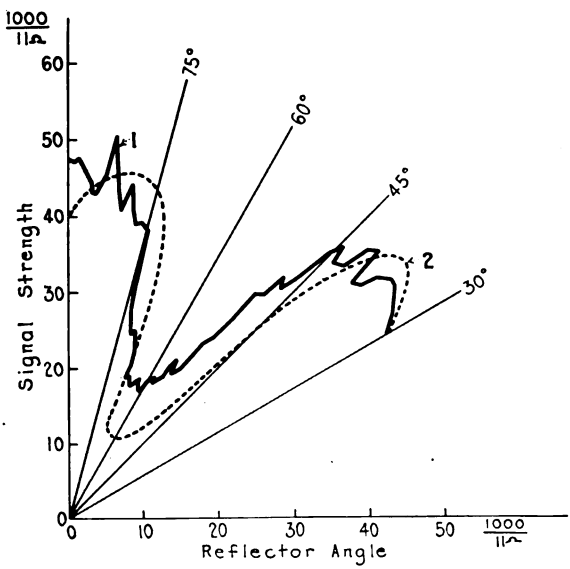

Figure 7—Signal strength for variation of reflector angle.

Mr. A. Ellerbrock in Rio de Janeiro and later in Buenos Aires). As the reflector turned, observations in Rio de Janeiro indicated a maximum with the reflector at an angle of about 38 deg. A second maximum, usually 10 percent less, was observed at about 80 deg. In between lay a deep minimum. Fig. 6 shows a few typical reception curves. The turning of the reflector took four minutes. While the first three curves show definite maxima, the last curve does not. This lack of definite maxima occurred in about 10 per cent of the observations. There was no difference between forenoon and afternoon. In Fig. 7 the mean value of the field intensity is plotted for the various angular settings. The investigation was carried on in such a way that at first measurements were made for several days and mean values obtained for each setting of the reflector (mean value Fig. 6, Curve 2). Later

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measurements were made for several days in which the reflector was rotated between 30 deg. and 90 deg. during each four minutes. (Mean value for three days, Fig. 6, Curve 1). Frequent comparisons were made with a simple vertical antenna (3/4 wavelength long). The vertical antenna was seldom heard and when it was heard, the intensity was only a fraction of that obtained from the reflector (approximately 1/10).

From these results one would assume that on all other wavelengths as well as on 11 meters, an optimum radiation angle exists for transmission in which space radiation plays a part. At present investigations are in progress to determine this angle for 20 meters.

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MAKING NORMAL COORDINATES COINCIDE WITH THE MESHES OF AN ELECTRICAL NETWORK*

By

E. A. GUILLEMIN

(Massachusetts Institute of Technology, Cambridge, Massachusetts)

The theory of normal coordinates is perhaps best known and certainly most effectively interpreted physically in connection with a mechanical system of elastically-coupled mass points which carry out small oscillations about a point of equilibrium. Here a normal coordinate is such a direction in space upon which the projection of the oscillation of any point of the system contains only one of the natural frequencies of the system. There are as many such normal coordinates as there are normal frequencies; and since these directions are in general neither normal to each other nor coincide with the axes of the fundamental frame of reference, each has projections upon all coordinates of the reference system. Mathematically this is evidenced by the fact that the component oscillation of any mass-point of the system upon any reference coordinate contains in general all of the natural frequencies of the system. Furthermore, we can easily visualize a process of rotation of the reference coordinates which will bring at least one of them to coincide with a normal direction or normal coordinate. And when carried out mathematically, such a transformation will cause the normal frequency, belonging to the particular normal coordinate in question, to be confined entirely to the reference coordinate which has been brought into coincidence. The projected oscillations upon the remaining reference coordinates then lack this normal frequency, i.e., it has been eliminated from the rest of our reference coordinates.

In setting up the differential equations for the normal mode of oscillation of an electrical network, we find an exact mathematical analogy to those of a mechanical system having the same number of degrees of freedom. Hence, from a mathematical standpoint at least, there also exist in the electrical system certain normal coordinates. The physical interpretation of them is, however, somewhat more puzzling than in the case of the mechanical system, and it is the object of this paper to point out in as direct a way as possible what, in the electrical network, is the equivalent of the normal coordinate of a mechanical system; and how it is

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possible, by properly fixing the circuit constants, to bring about coincidence with certain meshes of the network, thereby confining the corresponding natural frequencies to these meshes and eliminating them from the rest.

We will begin with a brief review of the solution to the homogeneous system of equations describing the free oscillations of an electrical network. If we number the meshes of our system from 1 to n, and let $\lambda_{ii}, \rho_{ii}, \sigma_{ii}$ represent the total inductance, resistance, and elastance respectively present in mesh i, and further let $-\lambda_{ik}, -\rho_{ik}, -\sigma_{ik}$ represent the total inductance (here the algebraic sum of mutual and common self-inductance), resistance and elastance respectively common to meshes i and k, then obviously the countervoltage in mesh k due to current i_k in it is given by:

$$\lambda_{kk}\frac{di_k}{dt}+\rho_{kk}i_k+\sigma_{kk}\int i_k dt,$$

and the countervoltage induced into mesh k from mesh i will be given by:

$$\lambda_{ik}\frac{di_i}{dt}+\rho_{ik}i_i+\sigma_{ik}\int i_i dt.$$

In order to bring the analogy to the equivalent mechanical system still closer, we define:

$$x_k=\int i_k dt$$

as the corresponding "mesh charge," it being the equivalent of displacement in the mechanical sense just as electric current is the equivalent of velocity in the mechanical system. If we further use the abbreviation:

$$a_{ik}=\lambda_{ik}\frac{d^2}{dt^2}+\rho_{ik}\frac{d}{dt}+\sigma_{ik} \tag{1}$$

then the homogeneous system of simultaneous linear differential equations, which are an expression of Kirchoffs' e.m.f. law for the force-free network, become:

$$\left.\begin{aligned} a_{11}x_1+a_{12}x_2+\cdots+a_{1n}x_n&=0\\ a_{21}x_1+a_{22}x_2+\cdots+a_{2n}x_n&=0\\ \cdots\cdots\cdots\cdots\cdots\cdots\cdots&\\ a_{n1}x_1+a_{n2}x_2+\cdots+a_{nn}x_n&=0 \end{aligned}\right\} \tag{2}$$

We assume as solutions, the normal functions:

$$x_k=Y_k\epsilon^{pt} \tag{3}$$

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Substitution into (2) gives the algebraic system of condition equations:

$$\left.\begin{aligned} b_{11}Y_1+b_{12}Y_2+\cdots+b_{1n}Y_n&=0\\ b_{21}Y_1+b_{22}Y_2+\cdots+b_{2n}Y_n&=0\\ \cdots\cdots\cdots\cdots\cdots\cdots\cdots&\\ b_{n1}Y_1+b_{n2}Y_2+\cdots+b_{nn}Y_n&=0 \end{aligned}\right\} \tag{4}$$

where:

$$b_{ik}=\lambda_{ik}p^2+\rho_{ik}p+\sigma_{ik} \tag{5}$$

are now simple algebraic polynomials in p. If the conditions (4) can be satisfied in agreement with arbitrary initial conditions, then our assumed solution will be valid and useful. Obviously this eliminates the trivial solutions:

$$Y_1=Y_2=\cdots=Y_n=0.$$

Before going on with the solution in the usual manner, let us give the conditions (4) a more definite interpretation. In the first place it will be noted that the electrical equivalent to a physical dimension in the mechanical system is a mesh; and that the number of meshes in the electrical system is equivalent to the number of dimensions of the corresponding mechanical system. For the latter this number is limited to three in our actual space. In the electrical system, this number is physically unlimited—and fortunately mathematically also, the system (2) being written for an n-dimensional space. Now let us define the vector $\boldsymbol{b}_1$ with the components:

$$\boldsymbol{b}_{11},\ \boldsymbol{b}_{12},\ \cdots,\ \boldsymbol{b}_{1n}.$$

Similarly, the vector $\boldsymbol{b}_2$ with components:

$$\boldsymbol{b}_{21},\ \boldsymbol{b}_{22},\ \cdots,\ \boldsymbol{b}_{2n},$$

and so on down to a vector $\boldsymbol{b}_n$ with the components:

$$\boldsymbol{b}_{n1},\ \boldsymbol{b}_{n2},\ \cdots,\ \boldsymbol{b}_{nn},$$

the b_{ik}'s being those defined by (5). Then if we consider:

$$Y_1,\ Y_2\cdots Y_n$$

of equations (4) as components of a vector $\boldsymbol{Y}$ in our n-dimensional space, the condition equations (4) may be put in the vector form:

$$\left.\begin{aligned} (\boldsymbol{b}_1,\ \boldsymbol{Y})&=0\\ (\boldsymbol{b}_2,\ \boldsymbol{Y})&=0\\ \cdots\cdots\cdots&\\ (\boldsymbol{b}_n,\ \boldsymbol{Y})&=0 \end{aligned}\right\} \tag{6}$$

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where the round parentheses indicate scalar product. But if the scalar product of two vectors vanishes, it means that these are orthogonal to each other. Hence we see that the conditions (4), or their equivalent (6), demand that the vector $\boldsymbol{Y}$ be determined in such a way as to be *simultaneously orthogonal* to the given system of vectors $\boldsymbol{b}_1, \boldsymbol{b}_2, \cdots, \boldsymbol{b}_n$, assuming for the moment that the p's are known. But the components of $\boldsymbol{Y}$ we know to be the transient mesh charge amplitudes corresponding to the natural angular velocity p which fixes the $\boldsymbol{b}_{ik}$'s. Hence it is clear that the direction of our vector $\boldsymbol{Y}$ is the direction of the normal frequency. And we see at once that the condition equations (4) fix the directions of these normal coordinates.

We have so far said nothing about the possibility of fulfilling these condition equations. We shall give this a physical interpretation as well. Let us confine our argument for the moment to a three-dimensional problem so that the process of visualization becomes easier. Our coordinate axes are numbered 1, 2, 3 instead of being lettered x, y, z as is the usual case. Suppose then that the vectors $\boldsymbol{b}_i$ all have three components different from zero and are linearly independent. The conditions (4) or (6) require us to find such a direction in space as will be simultaneously orthogonal to all three of these vectors $\boldsymbol{b}_i$. Obviously this is impossible, since all the available dimensions are consumed by the given system $\boldsymbol{b}_i$. However, if the latter all lie in a plane, i.e., consume only two out of the available three dimensions, then the vector $\boldsymbol{Y}$ can be drawn normal to this plane to satisfy (4) or (6). And if the given system of vectors $\boldsymbol{b}_i$ all lie along the same direction, two independent solutions to (4) are possible. In general, if the given system of vectors $\boldsymbol{b}_i$ consume $(n-m)$ of the n available dimensions, then there are m possible independent solutions to the system (4). The determinant of the system is then said to have the rank $(n-m)$. If we demand a unique solution to the system (4), and this is the case with our network problem, then the rank of the determinant

$$D(p) = \begin{vmatrix} b_{11} \cdots b_{1n} \\ b_{21} \cdots b_{2n} \\ \cdots \\ b_{n1} \cdots b_{nn} \end{vmatrix} \tag{7}$$

should be $(n-1)$. This means that the determinant shall vanish, but that at least one of its first minors shall not vanish. The vanishing of the determinant:

$$D(p) = 0 \tag{8}$$

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is the well-known determinantal equation which fixes the natural angular velocities p.

The problem of making a normal coordinate coincide with one of the meshes of the given network is now clear. If for a given root $p=p_r$ of (8), an entire column in the determinant (7) vanishes, then obviously the normal coordinate will occupy the coordinate, i.e., mesh, thus vacated by the b_i's. Incidentally, if we make a column in (7) vanish, the corresponding row vanishes also due to the relation:

$$b_{ik}=b_{ki}.$$

To make a column vanish for one of the roots of (8) it is necessary that:

$$b_{i2}=\alpha b_{i1};\ b_{i3}=\beta b_{i1};\ \cdots\ b_{in}=\gamma b_{i1} \tag{9}$$

where $\alpha, \beta, \cdots, \gamma$ are any proportionality factors, not necessarily integers. For, expanding (7) by minors we have for instance for the first row:

$$D(p)=b_{11}B_{11}+b_{12}B_{12}+\cdots+b_{1n}B_{1n} \tag{10}$$

where the B_{ik} are the minors of b_{ik}. But by (9) we have for $i=1$:

$$D(p)=b_{11}\{B_{11}+\alpha B_{12}+\beta B_{13}+\cdots+\gamma B_{1n}\}. \tag{11}$$

The determinantal equation is factorable, and one of the natural frequencies is given by:

$$b_{11}=0. \tag{12}$$

It is this frequency which goes with the normal coordinate coinciding with mesh one in this case. Similarly we could have made any other mesh a normal coordinate. Mesh one in the above case will be the only one in which the frequency given by (12) can be present. The transient solutions for the other meshes will contain only $(n-1)$ terms. Obviously we can use this process to eliminate objectionable natural frequencies as will be shown later.

In general, mesh k becomes a normal coordinate when:

$$b_{ik}=\alpha_{ik}b_{kk} \tag{13}$$

where the α_{ik}'s are any proportionability constants. If (13) is satisfied for all the b_{ik}'s then all meshes become normal coordinates.

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The determinantal equation for this case becomes:

$$D(p)=(b_{11}\,b_{22}\cdots b_{nn})\begin{vmatrix}1 & \cdots & \cdots & \alpha_{1n}\\ \alpha_{21} & 1 & \cdots & \alpha_{2n}\\ \alpha_{31} & \cdots & 1 & \cdots & \alpha_{3n}\\ \cdots & \cdots & \cdots & \cdots\\ \alpha_{n1} & \cdots & \cdots & 1\end{vmatrix}=0. \quad (14)$$

But:
$$b_{ii}=\frac{\alpha_{ik}}{\alpha_{ki}}\,b_{kk},$$

so that (14) is equivalent to

$$b_{ii}=0. \quad (14a)$$

The entire network possesses only one natural frequency.

Suppose we have a primary mesh (mesh No. 1), coupled to $(n-1)$ other meshes and fix the circuit constants so that:

$$b_{1k}=\alpha_{1k}b_{kk}. \quad (15)$$

The other meshes are assumed not coupled to each other so that

$$b_{ik}=0 \text{ for } i\neq k\neq 1.$$

Then the determinantal equation becomes:

$$D(p)=\begin{vmatrix}b_{11} & \alpha_{12}b_{22} & \cdots & \alpha_{1n}b_{nn}\\ b_{21} & b_{22} & 0 & \cdots & 0\\ \cdots & \cdots & \cdots & \cdots & \cdots\\ b_{n1} & 0 & \cdots & b_{nn}\end{vmatrix}$$

$$=(b_{22}b_{33}\cdots b_{nn})\begin{vmatrix}b_{11} & \alpha_{12} & \cdots & \alpha_{1n}\\ b_{21} & 1 & 0 & \cdots & 0\\ \cdots & \cdots & \cdots & \cdots & \cdots\\ b_{n1} & 0 & \cdots & 1\end{vmatrix}=0. \quad (16)$$

The roots are obtained from:

$$b_{22}=0;\ b_{33}=0;\ \cdots b_{nn}=0; \quad (17)$$

and:

$$\begin{vmatrix}b_{11} & \alpha_{12} & \cdots & \alpha_{1n}\\ b_{21} & 1 & \cdots & 0\\ \cdots & \cdots & \cdots & \cdots\\ b_{n1} & 0 & \cdots & 1\end{vmatrix}=0. \quad (18)$$

All meshes except mesh No. 1 are normal coordinates, and the frequencies given by (17) are confined to their respective meshes. Mesh No. 1 contains only the one frequency defined by (18), which

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appears also in all the other meshes, since mesh No. 1 is the only one which is not a normal coordinate. If we made mesh No. 1 a normal coordinate also, the case would revert to the previous one, and all normal frequencies would coincide as given by (14a). In our present case, however, we have present in the network n different natural frequencies, but $(n-1)$ of these are pigeon-holed in their respective normal meshes and cannot appear in the rest of the network in the form of free or transient oscillations. When the network is subjected to suddenly applied forces in all of the meshes, the transient current in mesh No. 1 will contain only one frequency, while those in the other meshes will contain this one and also their own normal frequency as given by (17).

An interesting point in connection with the transient solution can be brought out if we consider a network where only one of the meshes, say mesh No. 1, is made a normal coordinate. The pair of conjugate complex roots, which define the one frequency in this normal coordinate, are given by:

$$b_{11}=0.$$

Let us indicate this root by $p=p_1$, and evaluate the corresponding transient current amplitude by Heaviside's formula. In the notation used in this paper, this formula becomes, for suddenly applied d-c. voltages:

$$i_{ik}=E\sum_{\nu}\frac{B_{ik}(p_\nu)\epsilon^{p_\nu t}}{\left(\frac{dD}{dp}\right)_{p=p_\nu}} \tag{19}$$

where the summation is to extend over all the roots of:

$$D(p)=0.$$

the subscripts ik on the current indicate that the latter belongs to mesh i with the voltage impressed in mesh k or vice versa. Suppose we assume the voltage impressed in mesh No. 1. Then the one term of (19) for $p=p_1$ becomes:

$$(i_{11})_{p=p_1}=E\frac{B_{11}(p_1)\epsilon^{p_1 t}}{\left(\frac{dD}{dp}\right)_{p=p_1}}. \tag{20}$$

But by the rule for differentiating determinants we have:

$$\left(\frac{dD}{dp}\right)_{p=p_1}=\left(\sum_1^n ik\frac{\partial D}{\partial b_{ik}}\frac{db_{ik}}{dp}\right)_{p=p_1}=\left(\sum_1^n ikB_{ik}(2\lambda_{ik}p+\rho_{ik})\right)_{p=p_1}$$

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But since for $p=p_1$ the first row and column of $D(p)$ vanish, we see that:

$$B_{ik}(p_1)=0 \text{ for } i=k\neq 1, \tag{21}$$

so that:

$$\left(\frac{dD}{dp}\right)_{p=p_1}=B_{11}(p_1)\cdot(2\lambda_{11}p+\rho_{11}),$$

and substituting into (20) we get:

$$(i_{11})_{p=p_1}=\frac{E\epsilon^{p_1 t}}{2\lambda_{11}p_1+\rho_{11}} \tag{22}$$

which shows us that the amplitude of this normal oscillation, which is confined to mesh No. 1, depends only upon the constants of this mesh and not at all upon those of the rest of the network. That the frequency given by $p=p_1$ is actually confined to mesh No. 1, is immediately evident from (21) and (19) and needs no further elaboration.

Suppose now that we impress our voltage in any of the other meshes, and proceed to calculate the transient current amplitude for $p=p_1$ by formula (19). Here again, due to (21) we find that this amplitude vanishes, even for mesh No. 1. Hence we note the very interesting fact that, not only is the normal frequency $p=p_1$ confined to the mesh which is its normal coordinate, but it will not come into action unless the actuating force is applied in that mesh. In the case where all meshes except one are normal coordinates, the transients in all meshes, due to a force applied in the non-normal mesh, contain only one term in spite of the fact that the whole network possesses n natural frequencies. In other words, a normal frequency which coincides with a mesh can only be actuated by a force applied in that mesh, and does not respond to disturbances which occur in the other meshes. Obviously the complexity of transients in complicated networks may be reduced considerably by making use of this fact.

The point of chief interest which this analysis has in connection with steady state operation is the elimination or reduction of the number of resonance maxima when the network is subjected to sustained variable frequencies. In simple coupled radio circuits containing only two meshes, the double resonance peak in the secondary is an objectionable feature and gives rise to "broad" waves. By making the primary a normal coordinate, one of these

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can be confined there, and the secondary will then contain only one resonance peak. Or by making both meshes normal coordinates, the network will contain only one frequency and hence only one resonance peak in primary or secondary. The second case would, however, be harder to maintain under variable conditions of secondary tuning and radiation;—but preference here is merely a matter of design.

In general, if a steady state e.m.f. of the form:

$$e(t) = Re\{E\epsilon^{\mu t}\} \tag{23}$$

(where μ may be real, imaginary, or complex, and *Re* denotes "real portion of") is impressed in mesh k, the resulting steady state charge in mesh i is given by:

$$x_{ik} = Re\left\{\frac{EB_{ik}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\}, \tag{24}$$

and hence the corresponding mesh current is given by:

$$i_{ik} = \dot{x}_{ik} = Re\left\{\frac{E\mu B_{ik}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\} \tag{25}$$

Again let mesh No. 1 become a normal coordinate, and let its normal frequency be represented by:

$$p_1 = -\delta \pm j\omega \tag{26}$$

Then for $i = k = 1$, and $\mu = j\omega$ we see that:

$$\left[\frac{B_{11}(\mu)}{D(\mu)}\right] \to \infty \text{ for } \rho_{ik} \to 0, \tag{27}$$

and hence it is apparent that mesh No. 1 resonates at this point. However, keeping the e.m.f. in mesh No. 1 and calculating the current in any other mesh we have:

$$i_{i1} = Re\left\{\frac{E\mu B_{i1}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\}. \tag{28}$$

But

$$\left[\frac{B_{i1}(\mu)}{D(\mu)}\right]_{i \neq 1} \xrightarrow{0} 0 \text{ for } \rho_{ik} \to 0. \tag{29}$$

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We can get at the behavior of this indeterminate form for large values of μ by evaluating in the usual manner. For the denominator we have:

$$\left[\sum_1^n ik \frac{\partial D}{\partial b_{ik}} \frac{db_{ik}}{d\rho_{ik}}\right]_{\rho_{ik}=0} = \sum_1^n ikB_{ik}^* \cdot \mu$$

where the * indicates the limit for $\rho_{ik}=0$. But again:

$$B_{ik}^* = 0 \text{ for } i \neq k \neq 1,$$

so that the coefficient of the linear term in the Taylor expansion of the denominator becomes:

$$B_{11}^* \cdot \mu.$$

For the numerator we have:

$$\left[\sum_1^n rs \frac{\partial B_{i1}}{\partial b_{rs}} \frac{db_{rs}}{d\rho_{rs}}\right]_{\rho_{rs}=0} = \sum_1^{n\prime} rsB_{\{rs\}}^{\{i1\}} \cdot \mu \qquad \dagger$$

where the prime on the sum indicates that the terms for $r=i$ and $s=1$ are to be omitted. But here we see that:

$$B_{\{rs\}}^{*\{i1\}} = 0 \text{ for } r \neq 1,$$

so that the corresponding Taylor coefficient becomes:

$$\sum_2^n sB_{\{1s\}}^{*\{i1\}} \cdot \mu,$$

and our indeterminate (29) evaluates to:

$$\frac{\sum_2^n sB_{\{1s\}}^{*\{i1\}}}{B_{11}^*}. \qquad (29a)$$

Here the denominator is two powers higher in μ than the numerator, and hence the factor (29) varies as $1/\mu^2$ for large μ. Hence it is clear that all meshes except No. 1 do not resonate to the normal frequency p_1; and that for large values of the latter, the steady state oscillations in all but mesh No. 1 are very much suppressed. Mesh No. 1 tends to confine its normal frequency even in the steady state.

† The minor $B_{\{rs\}}^{\{ik\}}$ is obtained from $D(\mu)$ by striking the rows i and r and the columns k and s, and prefixing the sign $(-1)^{i+k+r+s}$.

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If now we put the source of e.m.f. in any other mesh except No. 1 and try to feed the normal frequency $\mu = j\omega$ into it, we have:

$$i_{1i} = Re\left\{\frac{E\mu B_{1i}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\}.$$

But due to the symmetry of $D(\mu)$, this is identical with (28), and hence we should arrive at (29a) again for the limit (29). In other words, it is just as hard to feed the resonance frequency into a normal mesh as it is to get it out of one. The normal mesh resonates to its own frequency only when the e.m.f. acts in it, and in that case tends to confine the oscillation there.

Obviously many other interesting combinations of normal and non-normal meshes may be gotten up and investigated both in the transient and steady states. The results of such investigations are often both interesting from the purely theoretical standpoint, as well as useful from the practical. It is the hope of the writer that the principles involved have been made sufficiently clear so that the reader may carry out his own extensions of the idea by himself.

Summary

The theory of normal coordinates in oscillatory systems is briefly reviewed, and their significance in connection with the electrical network pointed out. The relation which must hold between the circuit constants in order that normal coordinates may be made to coincide with certain meshes is derived. The effect of having made a certain mesh a normal coordinate of the system is then shown for the transient state, first, to confine the corresponding normal frequency to that mesh; secondly, to cause its amplitude to be independent of the other circuit constants; and, thirdly, to cause that normal frequency to remain inactive unless the actuating e.m.f. is impressed in that mesh. The effect so far as the steady state is concerned, is to confine resonance with the normal frequency in question to the normal mesh, and in general to present a means for eliminating or localizing resonance peaks in more complicated networks.

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VOLTAGE DETECTION COEFFICIENT*

By

E. L. Chaffee

(Cruft Laboratory, Harvard University)

In a previous paper[1] it was shown that the current of modulation frequency which flows in the plate circuit of a triode detector is given by the expression

$$\Delta^2 I_{pl} = (\boldsymbol{Det.\ I})_l \sqrt{2} m (\underline{\Delta E}_{oh})^2 \tag{1}$$

In the above equation and those that follow the Δ sign indicates small but finite values. Bold-face type indicates complex quantities. The subscript p refers to the plate of the triode, and the subscripts l and h refer to low or modulation and high or radio frequencies, respectively. m is the degree of modulation and $\underline{\Delta E}_{oh}$ is the maximum value (indicated by the underscore) of the unmodulated radio voltage impressed in the grid circuit of the triode detector. The notation used here and in the previous paper is fully described in a paper entitled "Vacuum Tube Nomenclature" published in these Proceedings by the author.[2]

The complex factor $(\boldsymbol{Det.\ I})_l$ is defined as the total current detection coefficient, but reference to the previous paper shows that the value of $(\boldsymbol{Det.\ I})_l$ depends not only upon the tube characteristics, but also upon the impedance of the particular plate load used during the measurement of $(\boldsymbol{Det.\ I})_l$. It is then not easy to determine the detection coefficient if the tube is used with some other plate impedance. If, however, the plate current $\Delta^2 I_{pl}$ be multiplied by the total plate circuit impedance a voltage is obtained which is a fictitious voltage but which, although non-existent, may be used for the calculation of the plate current of modulation frequency when the particular plate load is known. It was suggested in the previous paper[1] that a voltage detection coefficient may be defined according to the relation

$$\widehat{\Delta^2 E_l} = (\boldsymbol{Det.\ E})_l \sqrt{2} m (\underline{\Delta E}_{oh})^2 \tag{2}$$

which is more fundamental and a better quantity in terms of which to express the detection obtained with a particular triode.

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Paper presented before the American Section, International Union of Scientific Radio-Telegraphy, April 21, 1927.

[1] A Theoretical and Experimental Investigation of Detection for Small Signals by E. L. Chaffee and G. H. Browning, I.R.E. Feb. 1927, Vol. 15, No. 2.

[2] Vacuum Tube Nomenclature by E. L. Chaffee. I.R.E. March, 1927, Vol. 15, No 3.

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It is convenient to think of $(Det.\ E)_1$ as the fictitious voltage introduced into the plate circuit by a 70.7 o/o modulated signal of one-volt maximum amplitude if the detection followed the same law for so large a signal. It is probably better to consider one ten-thousandth of $(Det.\ E)_1$ as the equivalent audio voltage in the plate circuit due to a 70.7 o/o modulated signal of .01-volt maximum amplitude.

In the paper referred to above a method is given for measuring $\Delta^2 I_{p1}$ from which, knowing m and $(\underline{\Delta E}_{oh})$, the value of $(Det.\ I)_1$ can be calculated. The purpose of the present paper is to present a method of directly measuring $\widehat{\Delta^2 E_1}$ from which $(Det.\ E)_1$ can be calculated. Some experimental results are here presented to show the application of the method.

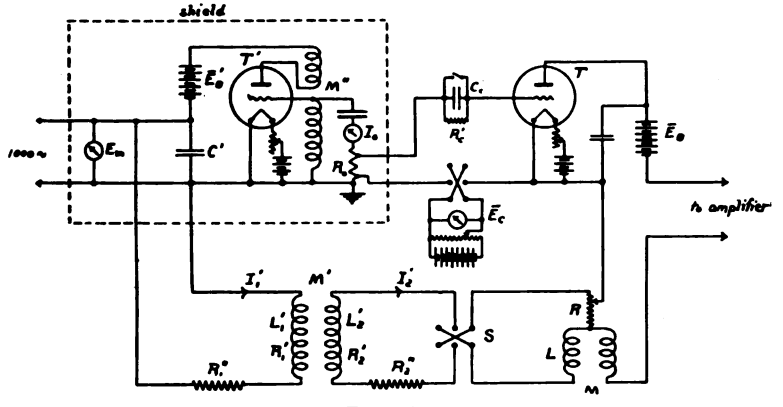

Figure 1

The diagram of connections for measuring the fictitious voltage $\widehat{\Delta^2 E_1}$ is shown in Fig. 1. Tube T is the triode under test. T' is a small radio-frequency oscillator, the output of which is modulated by an external source of power at 1,000 cycles. The oscillator is preferably enclosed in a shield shown by the dotted rectangle surrounding T' and its circuits. The oscillator has the oscillatory circuit connected to its grid. The plate circuit of the oscillator is regeneratively coupled to the oscillatory circuit through the mutual inductance M''. The plate voltage of the oscillator is pulsating and is obtained by adding to the steady voltage $\overline{E}_B'$, the modulating 1,000-cycle voltage E_m. The condenser C' serves only as a by-pass around the source E_m for the high-frequency currents in the plate circuit. The modulated oscillatory current passes through a small four-terminal resistance R_o, consisting of a short piece of

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fine resistance wire on which are soldered the potential taps. The voltage drop across the two potential terminals is introduced into the grid circuit of the tube T. Also in the grid circuit are the potential-changing means to provide polarizing potential $\overline{E}_c$, and the grid condenser and resistance C_c' and R_c'. The latter may be short-circuited when only plate-circuit rectification is considered.

Rectification taking place in tube T, in effect, introduces in the plate circuit a 1,000-cycle voltage $\Delta^2 E_1$. This voltage is opposed by an equal and opposite known voltage introduced into the plate circuit, balance being determined by silence in telephones connected to a two-stage amplifier. The known opposing voltage is obtained by passing through a variable resistance R and the coil L of the variable mutual inductance M a 1,000-cycle current derived from the modulating source E_m. An air-core, closely-coupled transformer, having mutual inductance M', is introduced and the circuit constants so adjusted that I_2' is in phase with E_m. To make possible any phase of the balancing voltage the current I_2' can be reversed by switch S, and M can be reversed by rotation of the secondary coil of M.

The condition under which the current I_2' is in phase with the modulating voltage E_m is simply stated below. Consider first, apart from the particular circuits of Fig. 1, a primary circuit which has a resistance R_1 and inductance L_1 coupled by a mutual inductance M to a secondary circuit having resistance R_2 and inductance L_2. Then the complex expression for the secondary current when an e.m.f. of E_1 volts is impressed in the primary circuit is

$$I_2 = \frac{jM\omega E_1}{(R_1+jL_1\omega)(R_2+jL_2\omega)+M^2\omega^2} \qquad (3)$$

$$= \frac{jM\omega E_1}{(R_1R_2-L_1L_2\omega^2+M^2\omega^2)+j(R_1L_2\omega+R_2L_1\omega)}$$

If the first bracket of the denominator is made equal to zero then the secondary current is in phase with the e.m.f. E_1 and given by the expression,

$$I_2 = \frac{E_1M}{R_1L_2+R_2L_1} \qquad (4)$$

Placing the first bracket of the denominator of (3) equal to zero gives as the condition necessary for no phase difference between E_1 and I_2—

$$R_1R_2 = L_1L_2\omega^2 - M^2\omega^2 \text{ or} \qquad (5)$$

$$\eta_1\eta_2 = 1-\tau^2$$

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where $\eta_1 = \frac{R_1}{L_1\omega}$, $\eta_2 = \frac{R_2}{L_2\omega}$ and $\tau = \frac{M}{\sqrt{L_1 L_2}}$.

Applying the above theory to the present case as given by the diagram of Fig. 1, the condition to be fulfilled in order that the current I_2' be in phase with E_m is, by equation (5)

$$(R_1'+R_1'')(R_2'+R_2''+R) = L_1'(L_2'+L)\omega^2 - M'^2\omega^2 \quad (6)$$

The value of the secondary current I_2' is then obtained by applying equation (4) to the present case.

$$I_2' = \frac{M'E_m}{(R_1'+R_1'')(L_2'+L)+(R_2'+R_2''+R)L_1'} \quad (7)$$

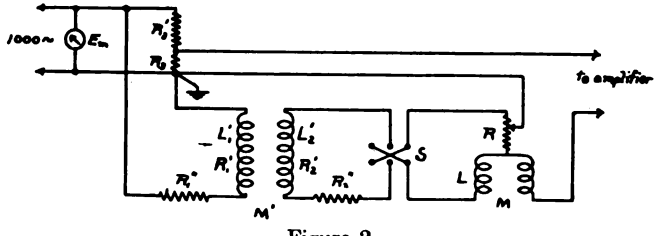

Figure 2

Although the proper values of the resistances of the circuit can be determined by calculation, usually it is easier to determine their proper values by experiment as follows: Referring to Fig. 2, two resistances, R_3 and R_3', are connected in series across E_m and the junction between the two resistances is connected to the terminal of the amplifier which in Fig. 1 is shown connected to the plate battery E_B. R_3 may conveniently be one-thousandth of R_2'. Then R_2'' and R_1'' can be so adjusted that with M equal to zero, balance is obtained with some value of R. The value of R gives the number of ohms which introduces into the plate circuit a voltage equal to $\frac{E_m R_3}{R_3+R_3'}$. It is usually desirable to make R_2'' about 10,000 ohms so that the variation of R in normal operation causes only a small percentage variation in the total secondary resistance. R is shown in the diagram as a potentiometer on the high-dial end. In the experiments to be described R was a three-dial box having dials of tenths, units and tens of ohms. As just stated, the connections were changed so that all of the 10-ohm coils were in the secondary circuit, the switch arm being represented

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by the arrow in the diagram. The total maximum change of resistance due to varying the two lower dials was then only 11 ohms.

Before measurements of detection coefficient can be made the oscillator T' must be adjusted to give sinusoidal modulation. The oscillator is tested by short-circuiting the alternating voltage E_m and reading I_o for various values of $\overline{E}_B'$. The value of coupling M'' and the grid polarizing potential of the oscillator should be so chosen that the plot of I_o against $\overline{E}_B'$ is straight over a considerable range of $\overline{E}_B'$, as shown in Fig. 3. A value of $\overline{E}_B'$ is then chosen corresponding to a point about midway on the straight portion of

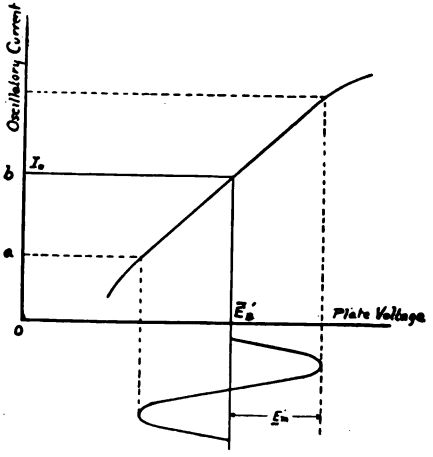

Figure 3

the plot. The peak value of E_m or $\underline{E}_m$ is then chosen to vary the plate voltage over the straight line portion of the plot. The degree of modulation m is then given by the ratio $\frac{\overline{ab}}{\overline{ob}}$ where $\overline{ob}$, as indicated in Fig. 3, is equal to I_o, the r.m.s. value of the unmodulated radio current, and $\overline{oa}$ is the minimum value of the r.m.s. radio current as it is modulated. It is assumed that the amplitude of the radio-frequency current changes as its plate voltage is varied at 1,000-cycles according to the plot of Fig. 3 which was obtained by a slow variation of plate voltage.

The value of $(\underline{\Delta E}_{oh})$ to be used in equation (2) is obviously

$$\underline{\Delta E}_{oh} = \sqrt{2} I_o R_o \tag{8}$$

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where I_o is the value of the radio current when there is no modulation. When the modulating e.m.f. E_m is superposed upon the steady plate-battery voltage $\overline{E}_B'$, the radio frequency current will increase to $I_o\sqrt{1+\frac{m^2}{2}}$.

Some tests were made upon two tubes of the same size and make, except one has an amplification factor of 8.5 and the other a μ of 20. The low-amplification-factor tube corresponds to the type 201-A Radiotron.

The constants of the circuits during the tests described in this communication are given below:

$\overline{E}_B' = 45$ volts	$M' = 1.36$ h.
$E_m = 10$ volts at 1000 cycles	$L_1' = 0.841$ h.
$m = 0.51$	$L_2' = 3.63$ h.
$I_o = 74.7$ milliamperes	$R_1' = 576.$ ohms
$R_o = 0.504$ ohms	$R_2' = 1490.$ ohms
$(\Delta E_{oh}) = 0.0533$ volts	$R_1'' = 3310$ ohms
	$R_2'' = 10000$ ohms

When calibrating the transformer by the method of Fig. 3 it was found that 87.5 ohms in R balanced $\frac{5}{1005}\times E_m$. Then

$$\widehat{\Delta^2 E} = \frac{5}{1005\times 87.5}\cdot\sqrt{R^2+M^2\omega^2}$$

$$= 0.000568\sqrt{R^2+M^2\omega^2} \qquad (9)$$

and $$Det.\ E = \frac{0.000568}{\sqrt{2}\times 0.51\times 0.0533^2}\cdot\sqrt{R^2+M^2\omega^2}$$

$$= 0.278\sqrt{R^2+M^2\omega^2}$$

The curves of Figs. 4 and 5 which exhibit observed points give the values of (*Det. E*) for plate-circuit rectification plotted against the grid polarizing potentials. The three curves of Fig. 4 are for the tube possessing an amplification factor of 8.5 and the two curves of Fig. 5 give the results for the high-μ tube. The values of (*Det. E*) increase as the plate current, plotted against grid volts, approaches zero due to the increasing resistance of the tube. The curves end when the current has become so small that balance is indefinite.

Not all of the audio voltage given by (*Det. E*) is available as

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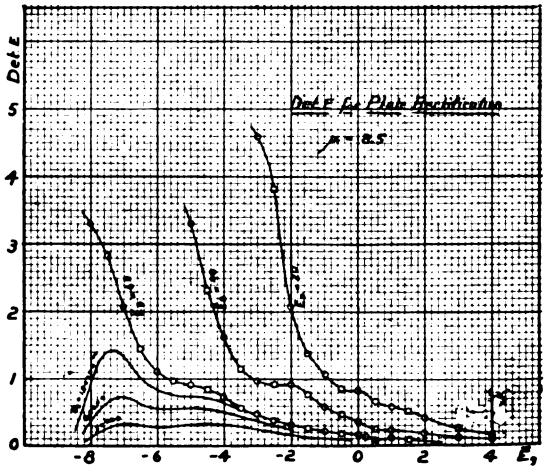

Figure 4

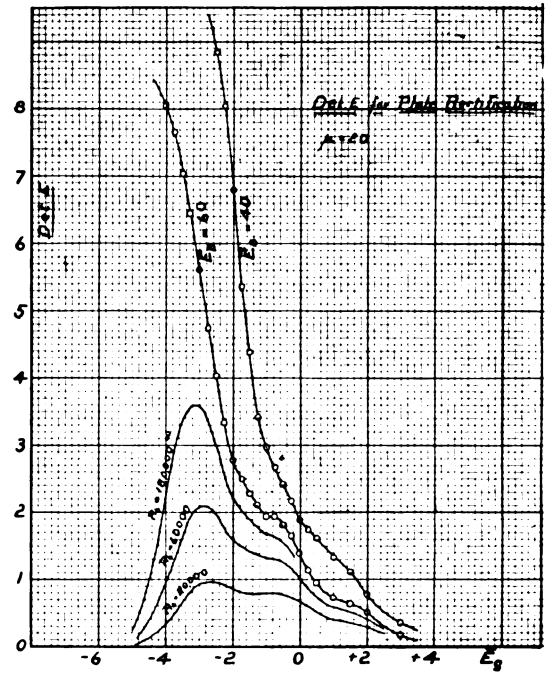

Figure 5

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output voltage to act upon the next amplifier tube. The available fraction of (*Det. E*) is obtained by multiplying by the ratio $\frac{\sqrt{R_b^2+X_b^2}}{\sqrt{(r_p+R_b)^2+X_b^2}}$, where R_b and X_b are the resistance and reactance of the load in the plate circuit of the detector. It is apparent then that, given the curves of (*Det. E*) and of r_p for a tube, the output voltage can be obtained for any plate load. If the plate load has no reactance then the ratio becomes $\frac{R_b}{r_p+R_b}$. The

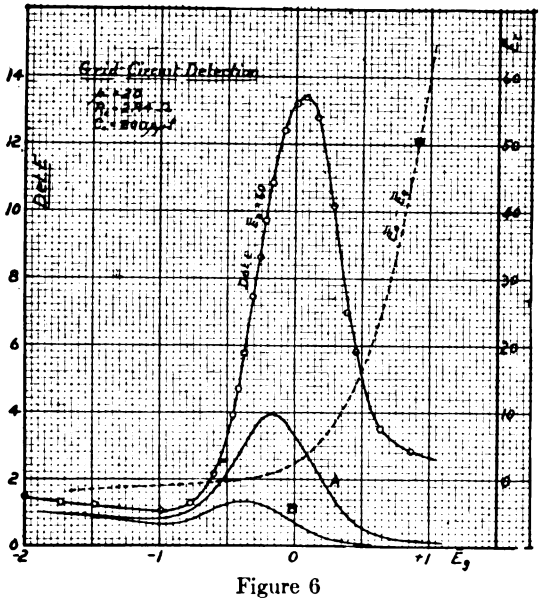

Figure 6

three full-line curves of Figs. 4 and 5 were obtained by multiplying the curve of (*Det. E*) for 60 volts by the above fraction for three values of R_b given on the curves. It is worthy of note that the values of k_p for the two tubes at the grid voltage which gives maximum detection are not very different. The value of k_p at -7 volts for the low-μ tube is 8.2×10^{-6} mhos. and for the high-μ tube at -3 volts is 9.9×10^{-6} mhos. Examination of the curves shows that the high-μ tube is the better detector when used to give plate-circuit rectification but in either case it is desirable to use a plate load of high impedance.

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The curves of Figs. 6 and 7 which show the observed points give the values of (*Det. E*) for grid-circuit rectification, when the grid resistance is about 3 megohms and the grid condenser 200 micro-microfarads. Curves of Fig. 6 are for the low-μ tube and those of Fig. 7 are for the high-μ tube, both for a plate potential of 60 volts. The curves are plotted against *actual* grid volts, but since the grid takes current when the grid voltage is positive there is a

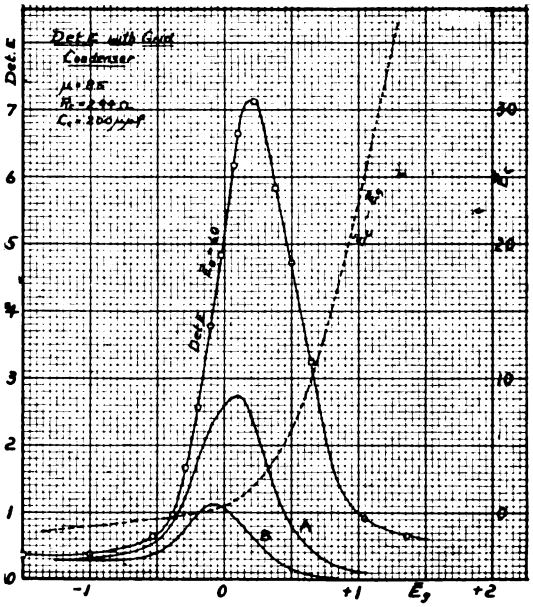

Figure 7

large voltage drop through the external grid resistance necessitating sometimes a very large applied polarizing potential $\overline{E}_c$ to obtain even a slightly positive value of $\overline{E}_g$. The dotted curves of Figs. 6 and 7 give the relations between $\overline{E}_g$ and $\overline{E}_c$.

In the case of grid-circuit rectification there are two effects which operate to cause the actual available voltage output to be less than that given by (*Det. E*). In the first place only a certain fraction of the fictitious voltage acting in the plate circuit is available as output voltage as explained in the case of plate-circuit rectification. The ratio $\frac{\sqrt{R_b^2+X_b^2}}{\sqrt{(r_p+R_b)^2+X_b^2)}}$ reduced to apply to a pure resistance plate load for the two cases shown in Figs. 6

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and 7, are plotted in Fig. 8. The full-line curves are for the low-μ tube; the dotted curves apply to the high-μ tube.

The second cause operating to reduce the output voltage in the case of grid-circuit rectification is the damping of the input oscillatory circuit due to the conduction between grid and filament. Let ΔE_{oh} be the radio voltage across the oscillatory circuit when there is no grid conduction as in plate-circuit rectification, and let

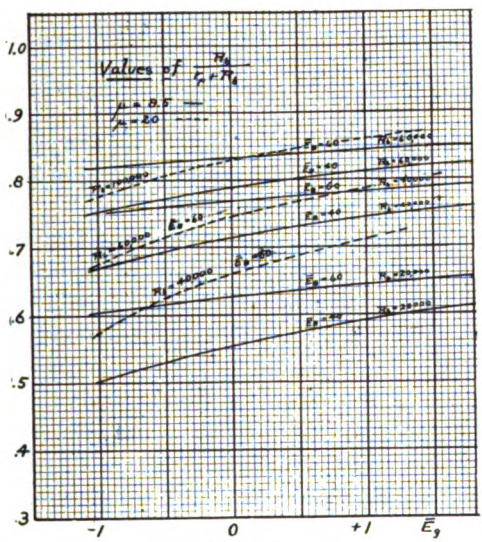

Figure 8

$\Delta E_{oh}'$ be the radio voltage across the same circuit when the grid conductance is different from zero and equal to k_g. The second reducing factor is then $\left(\frac{\Delta E_{oh}'}{\Delta E_{oh}}\right)^2$. The ratio is squared because detection varies as the square of the radio amplitude. It can be easily shown that this ratio has, to a sufficient accuracy, the value given in (10) below.

$$\left(\frac{\Delta E_{oh}'}{\Delta E_{oh}}\right)^2 = \left(\frac{\eta}{\eta + L\omega k_g}\right)^2 = \left(\frac{1}{1 + \frac{L^2\omega^2}{R} \cdot k_g}\right)^2 \qquad (10)$$

where η is the ratio $\frac{R}{L\omega}$ for the oscillatory circuit. The ratio $\frac{L^2\omega^2}{R}$ is the equivalent resistance of the oscillatory circuit at resonance.

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Referring again to Figs. 6 and 7, the full-line curves marked A were obtained from the detection-coefficient curves by multiplying by the product of the factor from equation (10) and the factor taken from Fig. 8 for the value of plate-load resistance R_b equal to five times the plate resistance r_p of the tube at zero grid volts. The values of R_b in the two cases are, respectively, 60,000 ohms and 100,000 ohms. For the two A-curves the value of $\frac{L^2\omega^2}{R}$ was taken as 100,000 ohms. The two curves marked B in Figs. 6 and 7 are similar to the A-curves except that the value of $\frac{L^2\omega^2}{R}$ was chosen to be 360,000 ohms. This value is a reasonable value because it corresponds to an η of 0.005, a value of L of 286 microhenrys, and a wavelength of 300 meters.

The enormous reduction in detection due to the factor of equation (10) is apparent. No account is taken of the effect of regeneration if it exists either solely because of the capacity between grid and plate of the tube, or if regeneration is purposely provided. Of course the effect of regeneration is to increase the voltage across the oscillatory circuit due to an e.m.f. induced in the coil, but such regeneration is probably more effective in increasing the signal strength with plate-circuit rectification than with grid-circuit rectification, for in the latter case, although regeneration decreases η, the factor of equation (10) is at the same time reduced. However, it is well known that since regeneration in both types of detection can bring the system to the point of oscillating, the signal strength then should be approximately the same in both cases. This is borne out by experimental observation.

The presence of the grid conduction acting across the oscillatory circuit very much broadens the resonance curve to the sacrifice of selectivity. The difference in selectivity in the two cases with no regeneration is very apparent in actual practice. In fact, taking into account all of the considerations, it seems to the author that grid-circuit rectification has little to justify its use either from the point of view of sensitivity or from the stand-point of quality.

Finally, it may be worth noting that the method of measurement outlined in the paper can easily be applied to the measurement of the voltages across a plate-load resistance or reactance, or across the secondary of a plate-circuit transformer, and in order to include the effects of damping of the oscillatory circuit by the

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grid conductance and to include the effects of natural or purposed regeneration, the radio voltage could conveniently be introduced in series with the inductance of an oscillatory circuit connected in the grid circuit of the detector. These changes might seem more satisfactory because the tube would be used as in practice. Such measurements are not here given because they are less definitive of the tube itself, but depend upon the particular impedances used.

Summary

The advantages of the use of voltage detection coefficient as a means of expressing the sensitivity of a detector and for the comparison of detectors are discussed. An experimental method is given for the measurement of the fictitious equivalent audio voltage which, if acting in the plate circuit of a detector, would give the demodulated audio plate current. Some experimental results are given for both plate-circuit detection and grid-circuit detection. The experimental results show that a high-μ tube is usually more sensitive as a detector than a low-μ tube.

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RADIO VISION*

By

C. Francis Jenkins

(Jenkins Laboratories, Washington, D. C.)

In speaking to you this evening on the subject of the electrical transmission of pictorial representations, may I say that in our laboratory we have found it convenient and informative to use the words radiogram, radiophone, and radio vision when we speak of radio-carried service; and to say telegram, telephone, or television whem we speak of wire-carried service.

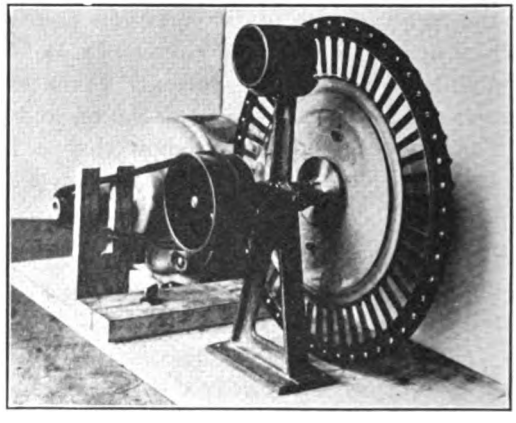

Figure 1—Laboratory receiver for radio vision. The black box contains a neon gas lamp; the slotted lens-disc sweeps the light-image spot across the screen in lines while the overlapping prism-disk distributes the lines from top to bottom of the screen.

The art of electrical picture-transmission is very old, relatively, for more than fifty years ago successful demonstrations were made in sending pictures by wire.

And there have been many workers, too, but the attainment of each was given but passing notice until the stamp of approval was put thereon by the great laboratories of the Bell Telephone Company, when, in April last, they made their spectacular demonstration between Washington and New York. This demonstration gave a great impetus to the development of electrical transmission of all kinds of pictorial representation.

* Original manuscript received by the Institute, July 6, 1927.
Presented before the Philadelphia Section of the Institute of Radio Engineers, June 24, 1927.

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In general there have been but two types of mechanisms employed, the cylinder and the flat surface. The cylinder has been used most often in the transmission of still pictures.

But obviously a flat receiving surface is the only possible type for radio vision and television, for the eye must seem to see the whole picture all the time, and that can only be done on a flat screen.

However, whatever type of machine is used, the only method employed to the present time consists in a linear analysis of the picture, scene, or object, and the instantaneous synthesis of each line on a distantly located receiving surface.

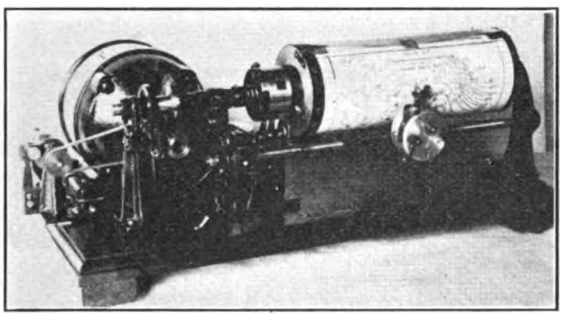

Figure 2—The weather map ship's receiver. The pen-box is moved across the base-map on the rotating cylinder by the screw, while the incoming radio signals touch the ink pen to the paper to build up the weather map.

The lights and darks of each successive line are changed into electrical current of corresponding strengths, which, carried to distant receivers, is there changed back into like light intensities and assembled on a suitable surface, for example, a sheet of paper, a photo film, or a flat picture screen.

It is quite evident that if this synthesis in the received picture is to result in an exact likeness to the transmitted subject, perfect synchronism of all receiving mechanisms with that of the transmitting mechanism must exist.

To attain synchronism, the synchronous motor is the simplest, but it is limited in application.

Other synchronizing methods have consisted of clock-controlled motors, where elapsed time is the standard against which all the motors were regulated.

Tuning forks for controlling the motors have been extensively employed, but this is but a modification, for it is only a smaller and more frequent division of time than the clock method.

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Because pendulum clocks and most tuning forks soon stop when installed aboard a rolling ship in a rough sea, an oscillator, resembling the escapement of a ship's chronometer, has been developed, and has been found quite a successful motor controller.

A simple method of synchronizing consists in maintaining the speed of the motor constant with a governor, together with means for automatically setting the receiving cylinder at zero on the beginning of each revolution of the transmitter cylinder.

We employed this method in the "three-cornered experiment" recently conducted by the U. S. Navy, the Weather Bureau, and the writer's laboratory.

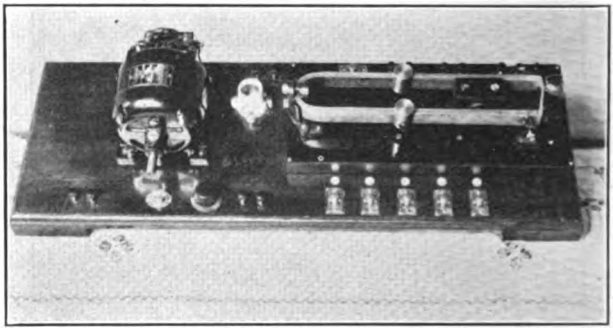

Figure 3—Fork-controlled motor unit. The vibration of the free ends of the fork arms cuts out armature resistance to keep the motor up to a definite speed. The fork continues to vibrate by reason of the current pulses in the electro-magnet between the fork arms.

Each morning the Weather Bureau made up a weather map, gave it to us, and we put it on our transmitter in the Navy building, which was connected by wire to the radio broadcast transmitters at Arlington, Virginia.

Weather map receivers were set up at certain land stations, and also on board the *U. S. S. Trenton*, and the *U. S. S. Kittery*. The latter made experimental cruises between Naval Operating Base at Hampton Roads and Caribbean Sea ports.

This territory was chosen for its well-known static disturbances, coincident with hurricane-forming zones. And it was found that weather maps could be recorded on the receivers with certainty even when weather information could not be received by code. The Florida hurricane of September provided a severe test of the system, de monstrating its worth and dependability.

While these "side-uses" are doubtless valuable, my premier ambition was radio vision, and so we do these other things only in "breathing spells" between attacks on the main problem.

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Our first public demonstration of radio vision occurred on June 13, 1925, when we showed in the laboratory in Washington, in the presence of Navy Secretary Wilbur, Admirals Taylor and Robinson, and many others, what was happening at the time at the Naval Air Station at Anacostia, some miles distant. It was the first radio vision demonstration ever made, I believe, and quite an historical event to many of us.

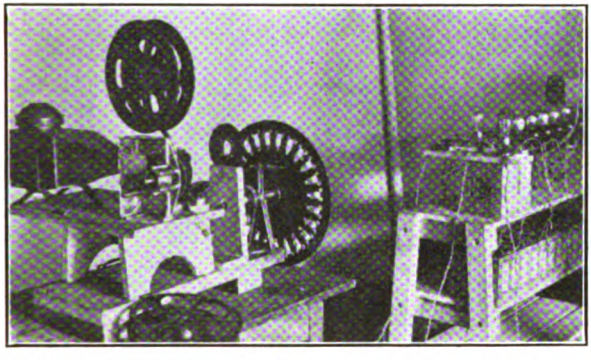

Figure 4—The simple mechanism by which motion pictures are transmitted. The rotation of the slotted lens-disk sweeps the image of each picture "frame" on the film across the light-sensitive cell in the box shown at the extreme right above the tubes.

The possibilities of radio vision for home entertainment, and of television in business, have been told so repeatedly in the public press that I hardly need restate here the promises of views of distant inaugural ceremonies, flower festivals, and baby parades. Being technical men I rather think you are more interested in details of methods and mechanisms.

As you know, the general scheme is to analyze the object or scene by a rotating scanning disk which in the transmitter permits the light reflected from the subject to fall on a light-sensitive cell, which, just as in still pictures, changes these light values into like current values; and which at the receiver permits light from a given source to be seen, directly or by reflection from a screen.

Such a scanning disk was shown in a patent as early as 1884. It consisted of a disk with one-fiftieth of an inch (1/50 in.) holes therein, arranged in a spiral.

The holes were an inch apart in the spiral, and the ends of the spiral had an inch offset; therefore, the contemplated picture was an inch square made up of fifty lines.

The light intensity, as in a pin-hole camera, is limited to the amount which can pass through this minute aperture, that is, 1/50 of 1/50, or only 1/2,500 part of the whole light.

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To overcome this limitation it was proposed by a Frenchman that an arc lamp of high-intensity be focused on these minute apertures, after passing which the spot of light is swept across the subject by the rapid rotation of the scanning disk. I hardly think

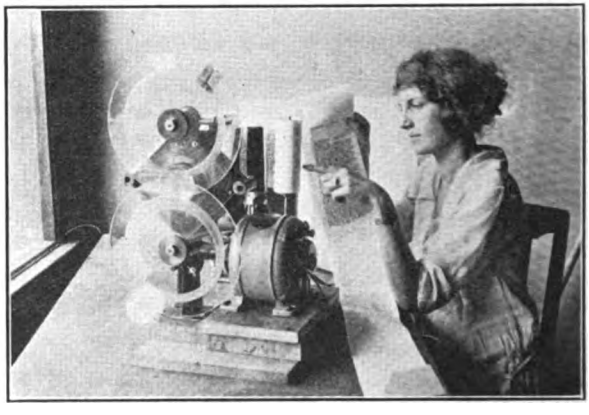

Figure 5—Strip device by which news copy, or like matter, can be transmitted by radio or by wire automatically as a continuous process. The overlapping prismatic rings in rotation sweep each typewritten line across a light-sensitive cell as the message moves longitudinally.

he expected to radio-transmit a baseball game by sweeping thereover a point of light from an arc lamp.

We, in the laboratory, think that limiting the light by passing it through these apertures is not the best plan; the available light is too limited. So, to get a greater value of the light, we usually make the openings in the scanning disk 1-1/2 inches in diameter and put lenses over the openings; and get the required tiny light-spot by focusing the light-source as a tiny flying spot on the receiving screen, to build up the moving picture.

While we have made a variety of mechanisms, this has been a fundamental principle in all of them.

Television would doubtless have been attained as early as the telephone if a suitable light-sensitive cell had been available for the transmitter, and an adequate light-source for the receiver.

Such tools are available today, and marked progress has been made in recent months toward a radio vision receiver acceptable to the public, and with the many hands and minds now engaged on the problem, I confidently believe its completion is the work of but a few months more.

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In the transmitter we need more sensitive light-cells, or perhaps I should say, light-cells giving greater current output, a current output which will more dependably start the first tube of an amplifier.

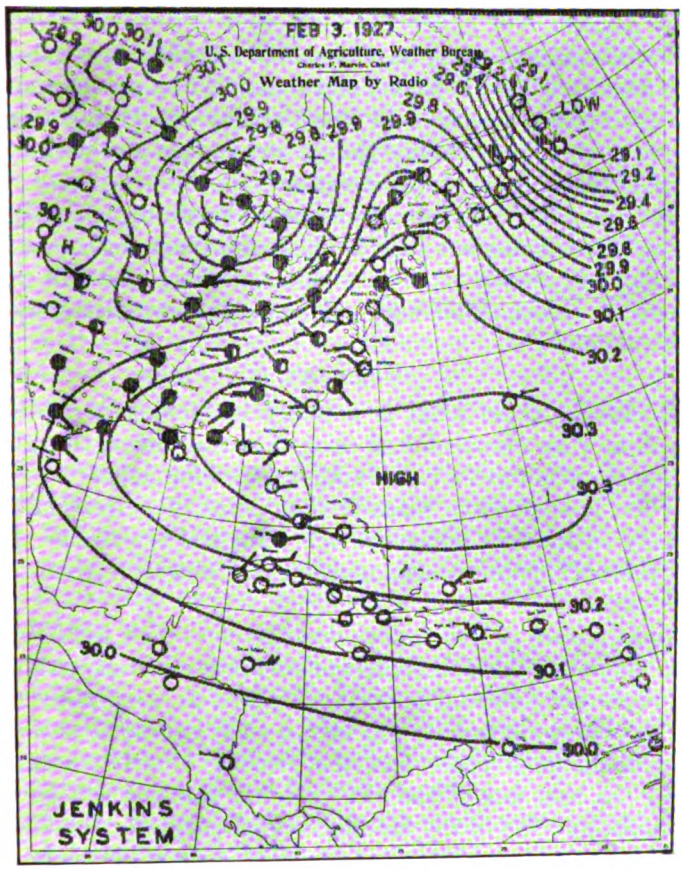

Figure 6—One of the weather maps as received aboard the *U. S. S. Kittery*. The base map is printed in brown and the isobars and other weather information received by radio is received thereon in red ink.

With light reflected from outdoor objects, the potassium cell gives but a very small current, only a few microamperes.

The present day potassium cells do not give current output proportional to the light intensity, but rather to the cell-surface covered by the light.

Resistance cells act too slowly for the required speed of light-reaction, which is of the order of 250,000 per second.

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Our problem in the receiver is a light-source more intense than the neon lamp, but having approximately its high-speed light-change.

The other possibility is a steady light source and a light-valve between the source and the screen for modulating the light to build up the picture.

For this purpose many have proposed a bisulphide cell through which polarized light is passed and controlled by a potential (the Kerr effect) or a current field (the Faraday effect), but this cell requires such a large energy change to produce any useful light change as to put it out of consideration.

Of course, we have our own proposed solutions, which I shall tell you about if they are successful. Radio vision for home entertainment will revive radio interest as nothing else will, for it combines the reach of the radio and the fascination of the story told in pantomime.

Summary

To see by radio what is actually happening at a distant place is now an accomplished scientific attainment, though the mechanism is not yet a merchandising development.

When it shall have reached that point of perfection, then one may sit in one's home and see inaugural ceremonies, baseball, football, polo games, mardi gras, flower festivals, and baby parades.

Radio vision is also experimentally combined with audible radio, and when these instruments are made generally available, then radio pictures and music and speech at the fireside, sent from distant world points, will be the daily source of news; the daily instructional class, and the evening entertainment; and equally the long day of the sick and the shut-ins will be more endurable, and life in the far places less lonely, for the flight of radio is not hindered by rain or storm, or snow blockades.

The electrical transmission of still pictures of photographs, sketches, maps, and the like, is now an every day affair, applicable to newspaper illustration and other usefulness.

But for the research worker it has lost its interest for him, it is too easy. To increase the speed of picture presentation ten thousand times, as required in radio movies, makes it a sporting proposition, and really worth while.

The apparatus emplayed in transmitting pen-and-ink sketches, photographs, and movies by radio are all simple in construction, though modified to best suit each particular use.

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BOOK REVIEWS

Radio Theory and Operating, BY MARY TEXANNA LOOMIS. LOOMIS PUBLISHING CO., THIRD EDITION, 886 Pages with approximately 700 illustrations, flexible covers. Price $3.50.

In the main this is a textbook for courses training students for commercial operating (although it might prove of value as a reference book for those engaged in other radio lines) and is written by the President of the Loomis Radio College.

A short historical section gives details of the experiments of Dr. Mahlon Loomis, of Washington, D. C., in 1865 et seq., who devised what is claimed to be a practical wireless telegraph system and to whom later a patent was granted. While definite proof and details of the system are unavailable, still many of the ideas indicated in the diagrams and notes reproduced anticipated modern practice and it would seem that Dr. Loomis should occupy a more important place in radio history than has been heretofore accorded him.

In common with many other textbooks it is used to best advantage in connection with lectures, since many of the terms used in the first few chapters are not adequately explained until later. Many cases will be noted where the use of definitions apparently from dictionaries do not disclose the exact distinctions usually accorded the terms.

A large number of inaccurate statements decrease the value of the book as a reference work. These are mainly confined to the first two hundred pages. These include such statements as: "A henry is defined as the energy which is induced by the cutting of one hundred million lines of force per second."

"A coil of high resistance wire passes all of the current in the circuit but consumes a portion of it in heat."

"The back e.m.f. due to inductance in a circuit is called reactance."

"These (referring to an illustration of a logarithmic condenser) are known as 'Straight Line' condensers, because when a graph is made of the capacity or of wavelengths in meters in a circuit in which they are used the 'curve' will be a straight line."

It is unusual that so many of these items should have survived to the third edition.

On the other hand, the major portion of the book is devoted to detailed description of apparatus circuits and procedure in broad-

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cast transmitting, commercial operating, amateur operating and receiving fields. Very little mathematical theory is included, but a number of useful tables are to be found.

The book is excellent guide for studying for a commercial operating examination. Pertinent questions and problems appear in each section, and a complete index of 27 pages completes the book.

R. R. BATCHER

Properties and Testing of Magnetic Materials, BY THOMAS SPOONER, McGRAW-HILL, 1927. FIRST EDITION, 385 Pages. Price $5.00.

The author of this book has admirably succeeded in gathering and arranging in useful form the more important properties of commercial ferromagnetic materials, and modern methods of testing and inspection. The book is singularly free of the accumulated "deadwood" of the early years of magnetic investigation, and although intended primarily for the use of the engineer it will also find a useful place in the library of the research worker.

GREENLEAF W. PICKARD

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued September 20, to October 11, 1927

BY

JOHN B. BRADY

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,642,688—FIXED CONDENSER—ALFRED MOSS, of New York, N. Y., Filed July 14, 1925, issued September 20, 1927. Assigned to Electrad, Inc.

1,642,861—ELECTRIC RELAY—L. B. TURNER, of Cambridge, England. Filed Feb. 13, 1919, issued Sept. 20, 1927. Assigned to Westinghouse Electric & Manufacturing Co.

1,643, 015—RADIO RECEIVING CIRCUITS— E. W. HOUSE, of Birmingham, Alabama. Filed August 27, 1925, issued Sept. 20, 1927.

1,643,323—DIRECTIVE ANTENNA ARRAY—JOHN STONE STONE of San Diego, Calif. Filed Jan. 4, 1921, issued Sept. 27, 1927. Assigned to American Telephone and Telegraph Co.

1,643,781—RADIO SENDING SYSTEM—DONALD G. LITTLE of Wilkinsburg, Pa., Filed Jan. 8, 1924, issued Sept. 27, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,643,782—DEVICE FOR ALTERING THE WAVE LENGTH—SIEGMUND LOEWE, Berlin, Germany. Filed Sept. 2, 1921, issued Sept. 27, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,644,242—CONDENSER—H. P. DONLE, Meriden, Conn. Filed Sept. 22, 1923, issued Oct. 4, 1927. Assigned to The Connecticut Telephone & Electric Co., Inc.

1,644,266—ANTENNA CONSTRUCTION—FREEMAN ORNE, Dubuque, Iowa. Filed Dec. 21, 1925, issued Oct. 4, 1927.

1,644,601—VACUUM DISCHARGE TUBE—A. P. HANS-GERD NICKEL, Charlottenburg, and JOHANNES J. SPANNIER, Berlin, Germany. Filed Feb. 24, 1927, issued Oct. 4, 1927.

1,644,744—ELECTRON TUBE—H. M. PINGEN, of Toledo, Ohio. Filed March 21, 1922, issued Oct. 11, 1927.

1,644,796—AMPLIFYING AND DETECTING TUBE—H. P. STUART, of Seattle, Washington. Filed March 27, 1926, issued Oct. 11, 1927.

1,644,906—FRAME AERIAL AND THE LIKE—P. W. WILLANS, of Pattishall, Towcester, England. Filed April 5, 1926, issued Oct. 11, 1927.

1,645,280—PHOTOELECTRIC CELL—T. R. GOLDSBOROUGH, of Wilkinsburg and OTTO H. ESCHHOLZ, of Pittsburgh, Pa. Filed Sept, 28, 1926, issued Oct. 11, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,645,291—POLYPHASE PLATE CIRCUIT EXCITATION SYSTEM—A. NYMAN, Swissvale, and FRANK CONRAD, of Pittsburgh, Pa. Filed August 18, 1921, issued Oct. 11, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,641,749—VARIABLE CONDENSER—JACOB M. ENDERS, Schenectady, N. Y. Filed Sept. 22, 1923, issued Sept. 6, 1927. Assigned to General Electric Company.

1,645,462—INSULATED SUPPORTS FOR INDUCTANCE COILS—LOUIS STEINBERGER of Brooklyn, N. Y., and GUY HILL, of Washington, D. C. Filed Dec. 11, 1922, issued Oct. 11, 1927. Assigned to Wired Radio, Inc.

1,645,231—ELECTROMAGNETIC SOUND REPRODUCERS—FREDERICK DIETRICH of N. Y. and WILLIAM H. GERNS of East Orange, N. J. Filed March 5, 1926, issued Oct. 11, 1927. Assigned to Brandes Laboratories, Inc.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED OCTOBER 5, 1927

Transferred to the Fellow grade

Connecticut,	Middletown	Cady, W. G.
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	New York City	Whittemore, L. E.
Pennsylvania,	Wilkinsburg	Batsel, M. C.

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New Zealand,	Auckland	Taylor, Conway

Elected to the Member grade

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New Zealand,	Auckland	Salt, G. McB.

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	Berkeley, 2529 Channing Way	Morgan, Howard K
	Los Angeles, 817 So. Parkview St.	Korntved, George W.
	Los Angeles, 207 So. Broadway	Nickels, Eugene E.
	Oakland, 2417 25th Avenue	Evans, George S.
	Santa Barbara, 22 E. Valerie St.	Means, Paul S.
	Santa Barbara, 1200 Anacapa St.	Sawyer, Gordon E.
	San Diego, USS S-44, c/o Postmaster	Wilcox, Nathan
	San Francisco, 30 Corona St.	Morris, John F.
Connecticut,	Bridgeport, 838 Park Avenue	Gould, Leslie
	Hartford, 29 Hartland St.	Kiley, Edward C.
	S. Norwalk, 91 N. Main St.	Nason, Charles H.
Dist. Columbia,	Washington, Navy Department	Cooper, Orville T.
	Washington, 1628 Riggs Place, N. W.	Kris, Joseph
	Washington, 226-8th Street, S. E.	Page, Robert M.
Florida,	Jacksonville, 2144 Rosselle St.	Winter, E. M.
Illinois,	Chicago, 950 Edgecomb Place	Allen, Justus
	Chicago, 6353 Yale Avenue	Beasley, Chas., Jr.
	Chicago, 510 Deming Place	Walter, V. W.
Indiana,	Fort Wayne, 314 Arcadia Ct.	Schryver, Henry A.
	Indianapolis, 1510 N. New Jersey St.	Ashlock, H. D.
Iowa	Cedar Rapids, 1834 Mollary St. West	Reschetrits, Hans.
Kentucky,	Newport, 520 East Second St.	Kolo, R. E.
Louisiana,	New Orleans, 1525 Eighth St.	Tcunisson, John F.
Massachusetts,	Boston, 316 Huntington Avenue	Gray, Alfred R.
	Cambridge, 4 Andover Hall, Harvard	Myers, G. B.
	N. Cambridge, 16 Allen Street	Gayne, Clifton
	Newtonville, 258 Linwood Avenue	Stark, Eldon E.
Michigan,	Flint, 1032½ Ann Arbor St.	Forbey, Wm. D.
Minnesota,	Elysian	Jarstad, E. O.
	Faribault	Carney, Philip G.
	Waseca	Johnson, Edgar F.
Missouri,	St. Louis, 4531 Emerson St.	Zehr, Nicholas J.
New Jersey,	Boonton, Radio Frequency Labs.	Haas, John G.
	Bound Brook, Station WJZ	Hagmann, Niklaus
	Bound Brook, Station WJZ	Lindstrom, A. J.
	Cliffwood, c/o Bell Telephone Labs.	Lowry, L. R.
	Elizabeth, 215 Inslee Place	Walsh, Lincoln
	Fort Monmouth, 51st Signal Battalion	Zimmerman, Julius
	Newark, 137 Fourth St.	Schiffenhaus, D. H.
	Vineland, 429 Elmer St.	Himes, Walter L.
New York,	Albany, 350 Second Ave.	Pozefsky, Leonard
	Brooklyn, 704 New Jersey Ave.	Ostrove, Wm.
	Brooklyn, 773 Willoughby Ave.	Kaplan, H. H.
	Buffalo, 1285 Jefferson Ave.	Barker, C. R.
	Buffalo, 95 Densmore Street	Baumann, N. C.
	Buffalo, 356 N. Ogden Street	Goets, Otto
	Buffalo, 18 Carmel Road	Hayslett, L. E.
	Buffalo, 215 Mulberry St.	Knowlton, E.W.
	Buffalo, 233 Sanders Road	Lawrence, A. P.
	Buffalo, 274 Delaware Ave.	March, Philip
	Buffalo, 25 Gelston St.	Miller, John G.
	Buffalo, 236 Barton St.	Miller, Theodore
	Buffalo, 42 Chatham Avenue	Schelling, R. F.
	Buffalo, 474 Carlton Street	Strauss, John B.

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	Buffalo, 38 Auchinvole Ave.	Voll, Harry
	Buffalo, 179 Royal Ave.	Watson, J. A.
	Hollis, L. I., 184–03 90th Ave.	Caputo, N. J.
	Jamaica, L. I., 8 George Street	McCord, Orison
	Maspette, L. I., 2573 Mymaud Place	Bennett, Jack
	New York City, 1416 Nelson Ave.	Alderman, L. T.
	New York City, 531 West 37th St.	Berry, James
	New York City, Van Cortland Park So.	Bunting, T. R.
	New York City, 241 West 108th St.	Horowitz, Saul
	New York City, 600 West 144th St.	Leach, H. B.
	New York City, 80 Cortland St	Schechter, Jack
	New York City, 144 West 72d St.	Van Blaricom, M.
	North Tonowanda, 84 Allen St.	Schurr, Wm.
	Port Washington, L. I., 69 Summet Road,	Hafner, G. M.
	Riverhead, L. I., Box 1191	Keary, H. F.
	Schenectady, General Electric Co.	Coombs, C. F.
	Schenectady, General Electric Co.	Hargrave, W. A.
	Schenectady, 209 Seward Place	Tittle, H. C.
	Woodside, 3740 65th St.	Bell, L. J.
	Yonkers, 579 Van Cortlandt and Park Ave.	Craig, A. G.
Ohio,	Alliance, 158 W. Cambridge St.	Dinger, H. E.
	Cincinnati, 234 West Fourth St.	Dixon, C. C.
Pennsylvania,	Bellevue, 233 South Starr Ave.	Donbar, C. F.
	Cresson, 301 Columbus Ave.	Meckley, R. C.
	Edgewood, 100 W. Hutchinson Ave.	Haak, Arthur
	Lester, 3d Ave. and Pontiac St.	Causse, R. D.
	Philadelphia, 1312 Arch St.	Stoudt, J. D.
	Philadelphia, 511 W. King St.	Trump, Wm. H.
	Pittsburgh, 1114 Gladys Ave.	Clark, C. M.
	Waynesburg	Kappes, A. J.
	Wilkinsburg, 1117 Wallace Ave.	Greis, W. K.
	Wilkinsburg, 193 Meade St.	Henderson, Roy
Texas,	Dallas, 1610 Commerce Street	Sternberg, F. B.
	Galveston, 1917 22nd Street	Hughes, K. E.
	San Antonio, 206 Laurel Heights Place	McShane, J. B.
Utah,	Murray, Box No. 48, R. D. No. 4	Smith, Edwin
Virginia,	Petersburg, 617 Union Ave.	Carr, R. J.
Washington,	Seattle, USS Jason, c/o Postmaster	Hurwitz, S. B.
Australia,	Melbourne, 44 Elizabeth St.	Grace, A.
	Victoria, 6 Esplanade Ave. Brighton Beach	Melhuish, C. A.
Canada,	Toronto, 217 Bowood Ave.	Hill, A. L.
	Ontario, 9 Gloucester St.	Mc Aleaney, H.
	Ontario, 805 Davenport Road	Parsons, J. B.
England,	Nr. Rugby, Cawston	Williams, Frank
	Middlesex, Northolt, Usebridge	Jamieson, J. R.
Japan,	Tokyo, Nippon Broadcasting Bldg.	Ueno, Shuzo

Elected to Junior grade

California,	Artesia, 438 N. Main St.	Corcoran, George
Connecticut,	Bridgeport, Sylvan Ave.	Ivanko, J. G.
Idaho,	Sandpoint, 422 Antone Street	Johnson, O. E.
Massachusetts,	East Pepperell, R. F. D. No. 1	Hills, E. K.
New York,	Buffalo, 145 High Street	Hoffman, Joseph
	Buffalo, 523 Massachusetts Ave	Hyman, Irving
	New York City, 600 W. 187th St.	Kambourian, Haig
	Rochester, 356 Seneca Parkway	De Witt, Wm. H.
	Rochester, 148 Colonial Road	Nelson, Edward

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RALPH BOWN
President, Institute of Radio Engineers—1927.

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 15 December, 1927 Number 12

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GENERAL INFORMATION

The PROCEEDINGS of the Institute are published monthly and contain the papers and the discussions thereon as presented at meetings.

Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

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INSTITUTE ACTIVITIES

November Meeting of the Board of Direction

On November 2nd the regular monthly meeting of the Board of Direction was held in the Offices of the Institute. The following were present: Ralph Bown, President; A. N. Goldsmith, Secretary; Melville Eastham, Treasurer; J. H. Dellinger, L. A. Hazeltine, R. A. Heising, J. V. L. Hogan, R. H. Manson, R. H. Marriott and J. M. Clayton, Assistant Secretary.

E. H. Armstrong was transferred to the grade of Fellow in the Institute. The following were transferred to the grade of Member: Wm. C. Ballard, Charles T. Burke, K. W. Jarvis, Charles E. Strong, W. G. Lush, and George T. Royden. The following were elected to the grade of Member: R. F. Durrant, Philip E. Glover and H. L. Roosevelt.

One hundred and twenty-seven Associate and eleven Junior members were elected.

1928 Convention Plans

The 1928 Convention of the Institute is to be held on January 9, 10, and 11. The program is to include the annual meeting of the Institute, reports of officers, installation of new officers, presentation of the Liebmann Memorial Prize, a series of technical sessions, inspection trips to a number of interesting places around New York and a dinner-dance on the last evening.

The Committees are actively at work on plans for the Convention. The membership of the Committee is as follows:

Convention Committee, Douglas Rigney, Chairman; J. V. L. Hogan, R. H. Marriott, G. C. Furness, R. H. Ranger, and W. G. H. Finch. Technical Sessions Committee: R. H. Marriott, Chairman; L. A. Hazeltine, A. A. Oswald, Paul Weeks, F. H. Kroger, Carl Dreher, Willis K. Wing, and A. Bailey. Trips Committee: R. H. Ranger, Chairman; O. B. Hanson, H. F. Dart, L. M. Clement, J. L. Reynolds, and N. Levinson. Dinner and Entertainment Committee: G. C. Furness, Chairman; G. W. Johnstone, Fulton Cutting, W. G. H. Finch, W. Wilson, H. T. Melhuish, and H. B. Coxhead. Publicity Committee: W. G. H. Finch, Chairman; D. G. Casem, O. E. Dunlap, F. J. F. Maher, Lloyd Jacquet, L. M. Cockaday, W. K. Wing, H. Gernsbach, L. W. Hatry, R. F. Yates, U. B. Ross, F. Ehlert, Austin Lescarboura, E.H. Hansen, J. G. Uzmann.

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A War Memorial to Engineers

No memorial has been created for American engineers who gave their lives in service outside this country during the world war, although nine years have passed since the Armistice. A unique opportunity for establishing a beautiful and acceptable memorial was found last June by the delegate sent to the five hundredth anniversary of the University of Louvain by the American Societies of Mechanical, Civil, Mining and Metallurgical, and Electrical Engineers. These Societies have accepted the opportunity and are initiating the plan of action. The Institute of Radio Engineers has been invited to participate.

The ancient library of Louvain was destroyed at the outbreak of the war. A beautiful new building is being erected with funds given wholly by Americans—almost a million of them—school children and teachers, college students and professors, policemen and firemen, and persons from all walks of life.

The new library has a fine tower in which space has been allotted for a clock and a carillon, but the clock and the carillon are the only items for which funds have not been provided. The privilege of giving them as a war memorial was secured for engineers in the United States. It is planned to have them ready for operation at the dedication of the building next May or June. Names of the engineers to be memorialized will be inscribed on tablets to be placed in the tower.

The Societies which have undertaken the project desire that the memorial shall embrace all American engineers who made the supreme sacrifice overseas, without limit to society membership, and that all American Engineers who can be reached may have the privilege of participating in this memorial. Contributions of five dollars or more are being requested, but any amounts which donors feel able to contribute will be quite acceptable. The names of all contributors, without mention of amounts, will be recorded in a handsome memorial volume to be deposited in the new Library.

Every dollar given will go into the memorial. Expenses are being provided for separately. A fund of $80,000 is needed to pay for the clock with four dials, the three-octave carillon with thirty-six bells, and to provide a sufficient fund for perpetual operation and maintenance. Contributions will be treated as a joint fund for all societies. The memorial will be given to the University of Louvain in the names of the participating societies.

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The Institute of Radio Engineers thinks that this is a very worthy project and feels that each member of the Institute will desire to contribute to the fund.

Contributions should be made payable to the *Committee on War Memorial to American Engineers*, and should be sent to the Institute's office with the form found on page xxiii of this issue of the PROCEEDINGS.

In order to complete the records of the Institute, it will be appreciated if each member will advise the Institute of the names of any members who died "over there" during the war. Space is provided on the form on page xxiii for inclusion of these names.

All contributions must reach the Institute before February 1, 1928.

RALPH BOWN, FIFTEENTH PRESIDENT OF THE INSTITUTE

Dr. Bown was born February 22, 1891 at Fairport, N. Y. After graduating from Fairport High School he went to Cornell University, where he obtained the M.E. degree in 1913. Following this he spent four years in the Department of Physics at Cornell as an instructor and graduate student, receiving the M.M.E. in 1915 and Ph. D. in 1917. During this period he became interested in radio and carried out some investigations on vacuum tube detectors. Going into the Army Signal Corps, Radio Division, first as a Lieutenant, he was in a few months promoted to Captain and became officer in charge of the radio development work at the Signal Corps Radio Laboratories at Camp Alfred Vail, N. J.

Upon discharge from the Army he became a member of the Department of Development and Research of the American Telephone and Telegraph Company and has been engaged in the various radio development activities of that company. One of his main interests has been endeavoring to develop the transmission side of radio engineering in a quantitative way. Together with his associates he has published a number of papers on various aspects of this work in connection with broadcasting and trans-Atlantic telephony.

Dr. Bown is a member of various scientific societies. He is a Fellow of the Institute of Radio Engineers and in 1925 was the chairman of its Committee on Standardization, in 1926 becoming vice-president of the Institute.

In 1919 he was married to Alma Crawford of Freehold, New Jersey. They have two sons. Dr. Bown's residence is at 85 Pine Street, Maplewood, N. J.

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Institute Meetings

New York Meeting

On the evening of November 2nd a paper by H. A. Wheeler on "Automatic Volume Control for Radio Receiving Sets" was presented to the New York meeting of the Institute by Mr. Wheeler. The meeting was held in the Engineering Societies Building.

In the discussion which followed the paper the following took part: G. W. Pickard, Lewis M. Hull, Edmund Bruce, H. A. Wheeler, L. A. Hazeltine, R. H. Marriott, A. N. Goldsmith, Ralph Batcher, Ralph Bown, and L. W. Hatry.

Canadian Section

At a meeting of the Canadian Section, held on October 5th in Room 23, Electrical Building, University of Toronto, W. R. McLachlan delivered a paper on "Broadcasting the Second Wrigley Marathon."

C. I. Soucy delivered the second of the series of Junior Lectures, the subject being "Condensers."

The attendance at this meeting was sixty.

On November 2nd a meeting of the Canadian Section was held in the Electrical Building, University of Toronto. C. T. Burke of the General Radio Company delivered a paper on "Radio Frequency Measurements."

Messrs. Price, Soucy, Smith, Hepburn and others took part in the discussion.

A. M. Patience delivered the third of the series of Junior Lectures entitled "The Oscillatory Circuit."

There were eighty-four persons in attendance.

The next meeting of the Section will be held on December 7th in the Electrical Building of the University of Toronto.

Chicago Section

A meeting of the Chicago Section was held on October 28th in the Monadnock Building. Montford Morrison presented a paper on "Methods of Reducing the Effects of Atmospheric Disturbances." Messrs. Minium, Harper, Wilcox and Henry discussed the paper.

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Cleveland Section

At a meeting of the Cleveland Section held on November 4, 1927 in the Case School of Applied Science, Cleveland, Dr. Ralph Bown delivered a paper entitled "Trans-Atlantic Telephony."

A general discussion followed the presentation of the paper.

The attendance at this meeting was sixty-four.

There will be another meeting of the Cleveland Section on December 2nd.

Connecticut Valley Section

On October 28th Dr. Ralph Bown delivered a talk on "Trans-Atlantic Telephony" before the Connecticut Valley Section of the Institute. The meeting was held in the assembly hall of the Traveler's Insurance Company, Hartford, Connecticut.

A general discussion followed the presentation of the talk.

Following the meeting, the members of the Institute in attendance (one hundred) visited the Travelers' Broadcasting Station WTIC.

Detroit Section

A meeting of the Detroit Section was held in the Conference Room, Detroit News Building, on October 21st. Professor Joseph Cannon of the University of Michigan delivered a paper on "Electric Filters."

There were forty-four members of the Section present.

Philadelphia Section

On October 28th a meeting of the Philadelphia Section was held in the Bartol Laboratories. C. Brown Hyatt delivered a paper entitled "Development and Application of Socket Power." Messrs. Nadick, Snyder, Tindell, Eaton, and Darlington discussed the paper.

There were fifty members of the Section present.

Committee Work

Sub-Committee on Power Supply

The Sub-Committee on Power Supply, of the Institute's Committee on Standardization, held a meeting at Institute Headquarters on November 3rd. Further progress was made on the proposed standard methods of ripple measurement and of measuring inductance of iron core inductors. Methods which have been developed and tried in some of the largest radio laboratories are be-

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ing scrutinized and tested by the committee members in an effort to reduce them to the simplest and most practical terms.

A list of terms and definitions applying to socket power units, rectifiers and filter systems was approved by unanimous vote of the committee for recommendation as new Institute standards.

Sub-Committee on Receiving Sets

A meeting of the Institute's Sub-Committee on Receiving Sets was held in New York on November 1st. Those present were: Dr. J. H. Dellinger, Chairman; E. E. Hiler, Professor L. A. Hazeltine, Professor C. A. Wright, W. A. Diehl, I. G. Maloff, L. C. F. Horle, Mr. Moore (representing V. M. Graham), E. T. Dickey (representing A. F. Van Dyck), W. M. Kirshbaum (representing M. C. Batsel) and M. Ferris (representing L. M. Hull).

The subcommittee considered the preliminary draft of its report as given in the printed pamphlet of May 20. Sections A, B, and C were approved with minor modifications. It was determined to add to Section D3 an alternative means of measuring field intensity by introducing voltage in a resistance in series with the antenna. It was also agreed to add a Section E, "Test Procedures." This Section is to give the details of procedure in carrying out the tests of sensitivity, selectivity, and fidelity. A Section F, to contain an improved bibliography, is also to be added. Subcommittees were appointed to prepare the additional material mentioned, and also to begin the drafting of test outlines of such additional set characteristics as distortion, radiation, and noise level.

A. E. S. C. Sectional Committee on Radio

Technical Committee on Vacuum Tubes

A meeting of this technical committee was held on November 2nd. Those present were: Dr. J. H. Dellinger, Chairman; Mr. W. C. White, Dr. M. J. Kelly, Dr. C. H. Sharp, Lieut. Comdr. W. J. Ruble and Mr. Stinchfield (representing R. M. Wise).

The Committee considered the standards previously recommended to the Sectional Committee on Radio. These standards included tube terminology and the UX base. After circulation by the Sectional Committee a number of comments of value had been received, and these received action by the technical committee.

Dr. Dellinger reported that the International Electro-technical Commission, at its meeting in Italy in September, had recommended for adoption two world standards for tube bases, [a]

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modified European base and the UX base. This action covered only the pin dimensions and spacing. The committee voted approval of the I. E. C. action, with an expression of opinion that international action may also be necessary on shell and bayonet pin dimensions.

The committee voted to adopt as the specific representation of the tube base standard the diagrams given on pages 55 and 56 of the September 1927 edition of the Natural Manufacturers Association "Handbook of Radio Standards." It was also decided to replace the designation UX by "American standard 4-pin vacuum tube base."

The Committee voted to submit to the consideration of the Institute of Radio Engineers Standardization Committee, the possible standardization of the measurement of tube capacities, the adoption of the term "tetrode" for 4-element vacuum tubes, and the possible standardization of terminology and other features of a.c. vacuum tubes.

The Committee determined not to undertake standardization at this time of filament voltage or current tolerances, or of internal capacity tolerances.

The Committee decided to undertake the assembly of data on the 5-pin base and on rectifier tube bases, with a view to possible future standardization.

1928 Member Dues

It is desirable that the attention of the membership be called to the bills for the 1928 dues which will be mailed during the latter part of December. Prompt payment of these bills will materially decrease the burden on the Institute office, and in addition will insure prompt receipt of all Institute publications by the membership.

Publication Schedule

Due to the publication of the Proceedings monthly since the first of the year and to a change in the printing establishment with the October number of the Proceedings, papers can be published in general in approximately one third the time formerly required.

The Institute will always appreciate constructive criticism on the Proceedings both as to the type of papers published and the mechanical make-up of the journal.

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1928 Year Book

Members are cautioned that changes in address (both residential and business) cannot be included in the 1928 Year Book unless the Institute is notified thereof by not later than December 15, 1927.

Unpaid members desiring to appear in the Year Book must settle their accounts on or before December 15th.

Membership Cards

As announced last month, membership cards for 1928 will be available to *all* members of the Institute when the Institute's office is requested to issue such cards.

Increase in Office Force

During the current year four additions have been made to the office staff of the Institute in an effort to keep up with the ever-increasing amount of work involved in the operation of the Institute and the publication of the Proceedings.

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OBITUARY

The Board of Direction of the Institute of Radio Engineers announces with deep sorrow the death of its member:

John F. Dillon

His career has been a notable one from the viewpoint of service to the public. For nearly thirty years he served in the Army of the United States, rising to the grade of Lieutenant-Colonel. For fifteen years he was a member of the Federal Radio Inspection Service of the United States Government, having been one of the first radio inspectors appointed in the United States.

At the height of his career he was appointed a member of the Federal Radio Commission in 1927. He served with the Commission in Washington in framing the preliminary policies and tentative regulations of that body until severe illness in June, 1927 prevented the continuance of his activities.

Colonel Dillon was an active worker in the radio field from its earliest days and was responsible for much of the development and coordination of radio communication on the Pacific Coast. He was a Fellow of the Institute of Radio Engineers, Chairman of the San Francisco Section of the Institute, and an active participant in Institute activities.

His loss is a real one to his country, the radio field, and the Institute, of which he was an officer.

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ATMOSPHERICS AT WATHEROO, WESTERN AUSTRALIA

By

J. E. I. Cairns

(Watheroo Magnetic Observatory, Western Australia. Carnegie Institution of Washington.)

Early in the year 1927 observations on the wave-form, field-strength, and duration of atmospherics were begun at the Watheroo Magnetic Observatory employing an apparatus essentially the same as that described by Appleton, Watson Watt, and Herd in the Proceedings of the Royal Society (*A*, Vol. 111, 1926), local conditions causing the introduction of only a few minor changes. For a month after the apparatus was set in proper working order, observations were made almost nightly, though no drawings were made, since at that time the calibrating oscillator for the time-base was in course of construction and rough calibrations only could be made by comparison of the sound produced in a pair of headphones with the note of a violin. The month was employed profitably by the observer's gaining experience in the delineation of the atmospheric wave-forms, and in increasing the speed with which the observations could be made. On March 18, 1927, the calibrating oscillator was delivered, and observations were begun on March 20.

It was first desired to measure 5,000 late summer atmospherics, but shortly after the first thousand had been drawn the oscillograph-tube became inoperative, owing either to a gas leak or to inactivation of the filament, and the observations had to be suspended. Summer atmospherics, especially late summer, are distinctly the worse experienced at this station, as elsewhere, and it was to measure these that the work was planned. Before the defective tube could be replaced the summer had broken, and consequently the statistical analysis has had to be made on 1000 wave-forms and not on 5000 as originally intended. However, it is believed that the result of this analysis is fairly typical. Several thousand atmospherics were examined during the month prior to March 20, 1927, and though some uncertainty attached to the values of the time-bases used, yet it may be stated that in both magnitude and duration, and in distribution as regards duration,

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these atmospherics were similar in every respect to those subjected to analysis.

Formulas

It is shown in the paper "On the Nature of Atmospherics," Parts II and III by Appleton, Watson Watt, and Herd,[1] that the effect of the rapid variation of the Earth's electric field which constitutes an atmospheric upon the system shown in Fig. 1 is given by

$$f(t) = \frac{v}{h} \cdot \frac{C_0 + C_1}{C_0} \tag{1}$$

where $f(t)$ is the change which the vertical electric field undergoes, h is the height of the exposed conductor, v is the potential developed

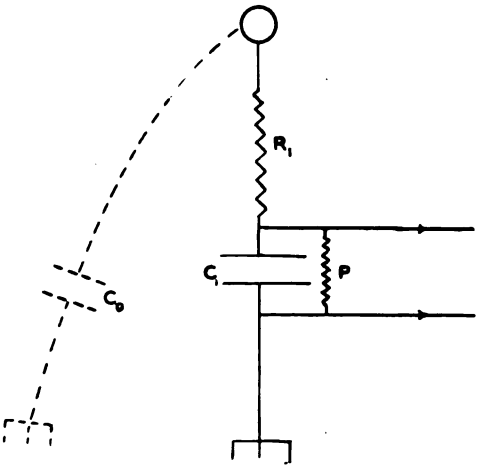

Figure 1

across the plates of the condenser C, and C_0 is the capacity of the exposed conductor, so that if the potential v be examined oscillographically it is possible, by means of the condenser potentiometer, to ascertain the changes in the Earth's field caused by an atmospheric for those cases where the initial and final values of the field are the same. For this to hold, the system comprised of C_0 and R_1 must be non-oscillatory and the time constant of the entire system must be small compared with the duration of the field change. Also, the time constant $C_1\rho$ must be large compared with the field change.

For a change involving a net change in the vertical field, a redistribution of charge on the exposed conductor takes place;

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at the end of the disturbance the charge induced on the conductor disappears by leakage through the resistance ρ in time $(C_0+C_1)\rho$. This exponential discharge produces a series of sloping lines on the oscillograph screen due to several tranverses of the indicating spot. The expression giving the maximum potential attained by the field change during a semi-permanent change is

$$[E]=\frac{v}{h}\cdot\frac{C_0+C_1}{C_0} \qquad (2)$$

where the symbols have the same significance as in (1).

Apparatus

The exposed conductor was an ordinary L receiving antenna of rather small dimensions; 91 meters long, 4.95 meters high at the center of capacity, and having a capacity of 0.00059 μf. With an inserted capacity C_1 of 0.002 μf. and an oscillograph sensitivity of 0.86 volts per mm., one mm. deflection on the oscillograph screen corresponded to a field-change of 0.00811 volts per meter, the amplification of the amplifier being 94. The resistance ρ was 12 megohms, and the lowest aerial resistance measured was 224 megohms, the insulation being maintained by means of sulphur strain insulators and a sulphur insulated lead-in tube. No measurements were made during wet weather, and when rain occurred, the insulators were scraped before the system was again used. Ground resistance was approximately 12 ohms.

The measuring apparatus is shown diagrammatically in Fig. 2, leads for the calibrating oscillator for the time-base and the time-base voltage generator itself being also shown. R_1, wound anti-inductively and anti-capacitatively, was just large enough so as to be greater than that required for critical damping, so as to give rapid response to field changes; ρ was a metallised non-inductive resistor. Switch S permitted the application of aerial voltage, amplified aerial voltage, or time-base calibrating oscillations as desired to the oscillograph plates. Otherwise the diagram is self-explanatory. The amplifier consisted of two stages of resistance capacity coupling, this form of amplifier being employed for reasons of economy, and also because the work was not of fundamental character. No change of amplification with frequency was observed for the range 256 to 1024 cycles, and the value of the

[1] E. V. Appleton, R. A. Watson Watt, and J. F. Herd, *Proc.* Roy. Soc. A, **3**. 1926.

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amplification was extrapolated backward to include the lower frequencies. The anode resistances, metallised, were continuously variable up to 80,000 ohms each, but none other than the maximum amplification was ever employed, the oscillograph deflections then being of convenient magnitude. The valves were Marconi Osram D. E. 5b, "high mu" valves, impedance 30,000 ohms. Rectification occurred in the valves when the grid potential exceeded 4 volts; when this value was attained at the grid of the second valve, the oscillograph spot was high on the

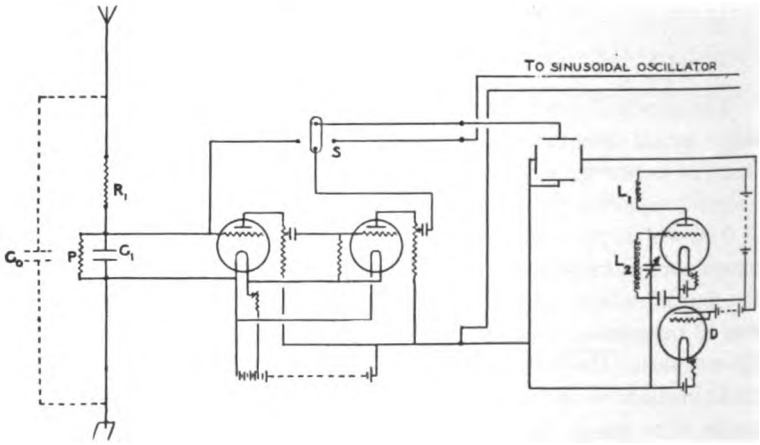

Figure 2—General Wiring Diagram.

screen, and on the portion not covered by the oscillograph calibration. Moreover, the image was confused, similar to a semi-permanent field-change with a radiation field superimposed. Such fields were readily recognised and omitted; no serious weighting of the results occurs as a result of neglecting these off-scale deflections, because of those deflections below 0.5 mm. which were also neglected. The omission at either end of the oscillograph range tends to preserve a correct mean.

The oscillograph was a standard Western Electric cathode-ray oscillograph, Vacuum Tube 224A. Anode potential used was 270, derived from radio plate-batteries, and filament current 1.4 amperes, the indicating spot being very brilliant at these values. All other vacuum tubes used in the apparatus were covered with light-proof sheaths to prevent excess stray illumination.

The audio-frequency oscillator employed to develop the sinusoidal oscillations for calibrating the time-bases was an

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ordinary valve generator, a Marconi Osram *R* valve being used for the purpose. The output voltage was only two volts and was therefore stepped up with a 5-to-1 a.f. transformer to give a larger oscillograph deflection.

A linear, uni-directional time-base was secured by the use of the non-linear oscillator described in the paper[1] previously mentioned. A short train of oscillations charges up the grid condenser until the oscillations cease; this charge then leaks off through the saturated diode *D* of Fig. 2 in a linear fashion until the voltage reaches the value where oscillations begin again. The application of the potential of the grid condenser plates during this process to the oscillograph plates produces a to-and-fro motion of the spot, the movement in one direction being far more rapid than in the other. The charging up of the grid condenser is practically instantaneous, so that the motion of the oscillograph spot is unobserved during this process. The coils L_1 and L_2 were ordinary tuning coils, suitably mounted, and it was found that the system oscillated better when the grid coil was tuned by a 0.0005-$\mu.f.$ variable condenser. It was also found necessary to screen this oscillator with an earthed metal screen; owing to the proximity of the time-base oscillator and the amplifier in the arrangement of the apparatus, there was considerable stray pick-up. When the earthed screen was introduced, the trace on the oscillograph screen with the aerial disconnected from the amplifier, was perfectly straight, whereas before it had shown a Lissajous figure. No trace of backstroke could be detected except at a frequency of 1024 p.p.s., when one wave of 512-cycle voltage was held on two strokes of the time-base; the return of the spot was just barely distinguishable, and as far as could be ascertained, parallel to the time-base.

Observations

The observations were made at night when weather conditions permitted. Usually at this time of the year at Watheroo the only rains are those due to thunderstorms, but unfortunately, heavy unseasonal rains from the northeast commenced almost simultaneously with the beginning of the observations and thus forced the observer to extend them over a longer period of time than would otherwise have been the case. Owing to the necessity of performing the ordinary observatory routine work, it was impossible to begin the observations until after 21^h; as these extended over three hours for the number of time-bases used and the period on each, there was danger of a diurnal variation creeping in and

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affecting the numerical distribution of the atmospherics with respect to their duration. To correct this effect, in the latter half of the series, the time-bases were used in the reverse order to the former: i.e. in the first half, time-bases of long duration were observed first, the duration being decreased as each time-base was examined, while in the second half, short-duration bases were

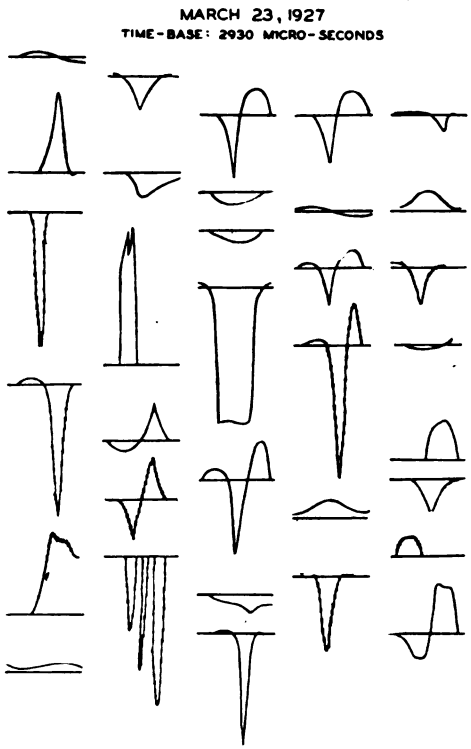

Figure 3—Observed Atmospherics.

observed first. This was the most satisfactory procedure, especially under humid conditions, for it was noticed that after rain had fallen, a slight leak across the grid condenser of the linear oscillator prevented the use of very low frequencies; when the short-duration bases were examined first, the apparatus became warmed up before these very long bases were examined, and no leak occurred. The ionisation in the oscillograph tube also provided a leakage path, but even at the lowest frequencies it did not appear to affect the linearity of the time-base. To offset the effect of this leak, (which is practically constant and linear), the largest grid condenser

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permissible was used in the time-base oscillator; in the actual apparatus it was 0.005 $\mu.f.$

Time-bases of 20-mm. length were used almost exclusively; on a few occasions 15-mm. bases were employed, but the thickening of the base-line caused their abandonment. Adjustment to length was effected by varying the coupling of the coils L_1 and L_2 of Fig. 2 of the time-base oscillator. The bases were calibrated against the sinusoidal oscillator which was itself standardised against a 512-cycle tuning fork by reducing the beat-note between the fork and the note produced in a loudspeaker connected to the sinusoidal

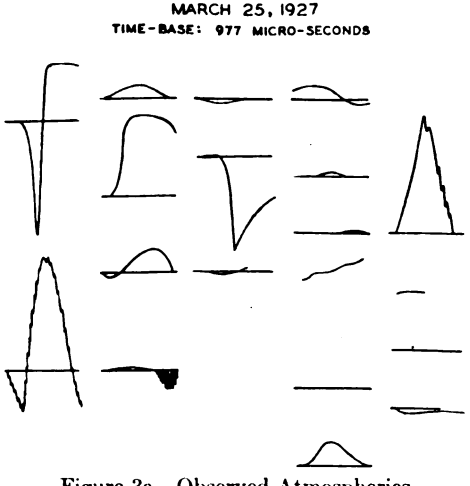

Figure 3a—Observed Atmospherics.

oscillator to zero. The time-base grid-condenser leak, the saturated diode D of Fig. 2, was then adjusted until the required pattern was held on the screen. Patterns containing up to six waves of the sinusoidal voltage could be held easily, and the more complicated patterns, such as where three waves were held on two strokes of the base, though a little difficult to interpret at first, were soon recognized with ease after a violin had been used to identify the audible note emitted. The strokes per second of the time-bases employed were as follows: 85.3, 102.4, 128.0, 170.7, 256.0, 341.3, 512.0, 768.0, and 1024 so that with bases of 20-mm. length, one mm. on these bases corresponded to 585.9, 488.3, 390.6, 293.0, 195.3, 146.5, 97.7, 65.1, and 48.8 micro-seconds respectively.

Originally it was intended to allot 15 minutes to observations on each base; one observer only made the observations and difficulty was experienced in keeping the observation time within the

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prescribed limits. In an attempt to correct for the differing durations of observation (amounting to ±3 minutes), the results have been weighted, though very little change in the unweighted figures is produced thereby.

Upon the end of the oscillograph tube, lines were marked parallel to the time-base and 5 mm. apart. Readings to 0.5 mm. could

TABLE I

SUMMARY OF ATMOSPHERICS AT WATHEROO MAGNETIC OBSERVATORY, WESTERN AUSTRALIA, MARCH 20 TO 28, 1927

Type[1]	No.	Duration in μs	Field strength in v/m	Type[1]	No.	Duration in μs	Field strength in v/m
Aperiodic Types, *P*							
A+	12	1793	0.153	*A*−	45	786	0.162
B+	36	1798	0.032	*B*−	54	1137	0.039
C+	56	3198	0.167	*C*−	71	2348	0.074
D+	9	2783	0.179	*D*−	21	1910	0.121
E+	21	1892	0.030	*E*−	44	1634	0.026
F+	21	1870	0.039	*F*−	16	1915	0.037
Indeterminate+	32	1553	0.095	Indeterminate−	33	2278	0.065
Total *P*+	187	2172	0.095	Total *P*−	284	1680	0.074
Quasi-periodic Types, *Q*							
Greater voltage first				Greater voltage first			
A+	21	3316	0.128	*A*−	29	3288	0.081
B+	10	3592	0.176	*B*−	1	4965	0.137
C+	15	2473	0.026	*C*−	15	2826	0.028
D+	0	....	.....	*D*−	0	....	.....
E+	3	4148	0.058	*E*−	0	....	.....
F+	0	....	.....	*F*−	0	....	.....
Indeterminate+	7	2415	0.035	Indeterminate−	0	....	.....
Greater voltage later				Greater voltage later			
G+	20	3227	0.186	*G*−	50	2410	0.144
H+	4	2478	0.148	*H*−	10	2576	0.190
I+	19	2448	0.051	*I*−	30	2722	0.058
J+	7	3014	0.123	*J*−	92	3458	0.132
K+	5	2538	0.081	*K*−	20	2788	0.065
L+	6	2749	0.199	*L*−	2	3738	0.234
M+	3	2606	0.220	*M*−	14	3176	0.232
N+	3	1531	0.269	*N*−	1	6095	0.202
O+	4	4320	0.227	*O*−	1	977	0.186
Indeterminate+	8	2076	0.107	Indeterminate−	31	2394	0.102
Total *Q*+	135	2876	0.118	Total *Q*−	296	2960	0.116
				F±	0	....	.....
A±	1	9020	0.026	*G*±	6	4410	0.019
B±	0	....	.....	*H*±	1	6975	0.021
C±	0	....	.....	Indeterminate±	16	1998	0.081
D±	0	....	.....				
E±	0	....	.....	Total *Q* ±	23	3126	0.060

[1] Type as defined in sketches of Fig. 5 either positive or negative.

be made without difficulty, so that the lowest value discernible corresponded to a field change of 0.004 volts per m. The time-base was adjusted to coincide with one of these parallel lines by means of the tuning condenser across L_2; change in the frequency of the oscillations constituting the short train which charged up the grid condenser appeared to affect the terminal voltages of the train

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and thus move the base as a whole parallel to itself. The adjustment was not very critical, but sufficient to correct any drift which occurred.

Owing to the necessity for the maintenance of high aerial insulation, special attention was given to this and the insulation tested twice nightly, before and after the observations, by means of a microammeter and a 200-volt battery. In no instance was the insulation resistance below 224 megohms, giving a minimum value of 11.4 megohms for the resistance ρ and therefore a value for the time constant $C_1\rho$ of 22,800 micro-seconds. In addition, a voltmeter could be inserted at various points to ensure the main-

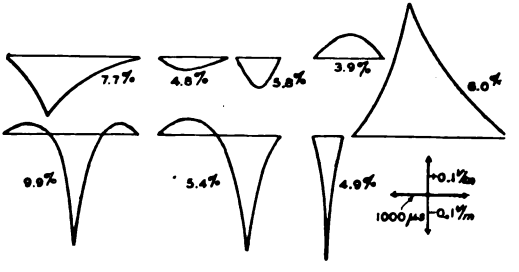

Figure 4—Reconstructed Typical Atmospherics.

tenance of the conditions necessary for the correct functioning of the amplifier and time-base oscillator.

Results

The wave-forms were drawn by hand; typical forms observed are shown in Fig. 3. In nearly all the atmospherics observed the structure was rippled; so much so that it was impossible to draw all the ripples, but merely to indicate on the drawing that ripple was present. When the time-base was opened out to 1000 μs these ripples were seen to be approximately 1 to 2 mm. in length corresponding in time to 50 to 100 μs. The results obtained are expressed concisely in Table 1, following the method of Appleton, Watson Watt, and Herd, and it will be noted that the wave-forms are identical with those obtained by these investigators. The curves in Fig. 4 are reconstructed from the mean values obtained for the most frequently occurring form in each class, and the individual curves may be taken as the typical atmospheric in that particular class. The percentages alongside each curve give the percentage with which the particular form occurred in the total number of radiation-fields examined (this excludes semi-

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permanent field-changes). The curves are not idealised to any great extent.

Negative atmospherics (the convention as to sign is that the atmospheric is given the sign of the greatest peak, regardless of the order of incidence of the peaks, and the positive sign is given to that atmospheric which tends to make the aerial terminal of the coupling condenser C_1 positive) predominate, though the peaked aperiodic positive forms an important class, comprising 6 per cent of the total number. These peaked positive aperiodics were almost invariably associated with visible lightning, and their magnitude is evidently due to the proximity of the source. The flashes were, in general, diffuse glows, but the district is so sparsely settled that no data as to distance could be obtained. Cloud-to-earth flashes are comparatively rare in the locality but "summer lightning" is frequently visible at night at the season of the observations. The symmetrical peaked negative is worthy of notice on account of the shortness of its duration, 786 μs. The peaked forms were far more numerous and of greater amplitude than the rounded, and the peaked quasi-periodics occurred more frequently, though attaining slightly smaller amplitudes on the whole, than the peaked aperiodics.

Rippled Structure

It was mentioned previously that ripple was present in nearly all the atmospherics examined; on nearly every occasion that observations were made the familiar summer "static roar" was present when a radio receiver was tuned to any station. It is believed that the cause of this "roar" is to be found in the small ripples of durations 50–100 μs. and of amplitude 0.008 volts per m. approximately from consideration of the following: Every night before commencing the observations, the apparatus was tested before darkness set in, and one or two stations were tuned in on the receiver. The static roar was always accompanied by ripples on the oscillograph screen when the time-base was opened out to 1000 μs., and on those occasions when the roar was absent (its absence was always marked) ripples were also absent. This coincidence occurred on several occasions in the 5 weeks during which observations were made, though only on one or two during the 9 days on which the most precise measurements were carried out. When the roar was very bad, these ripples were practically continuous; in fact, by varying the time-length of the base-line, it was possible on occasions to hold the pattern or to watch it

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progressing or retrogressing for a short period, a fact that led to the suspicion that local interference might be the cause. The electric motors driving other observatory apparatus were suspected, but shutting these off produced no change in the appearance on the oscillograph screen; in fact, all possible sources of stray pick-up were investigated but no evidence was found that there was any local interference. The only conclusion that can be reached is that the continuous summer roar is caused by continuous or almost continuous atmospheric ripples of from 50–100 μs. duration and about 0.008 volts per m. amplitude. No long-wave station employing a wavelength anywhere near the vicinity of

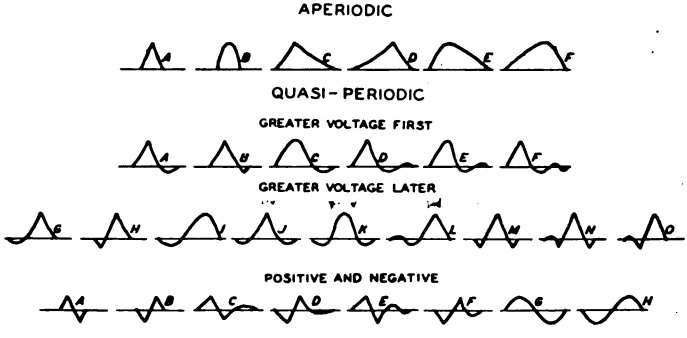

Figure 5—Types of Atmospherics.

13,330 metres (22.5 kc.) is situated in or near Australia, nor could such a station be tuned in on a receiver. Such interference is possible, because of the aperiodic aerial employed. This "roar" is most probably identical with the "rattle" which De Groot[2] observed in the Dutch East Indies, and which he ascribed to extra-terrestrial influence.

Semi-permanent Fields

Semi-permanent field-changes were fairly numerous, their incidence amounting to 17.6 per cent of the total number of 1124 atmospherics examined. The sign of a number of these was not noted on occasions, but of those which were measured the negative had a large predominance; positives were very few. One class, giving a complex screen pattern, was termed "±," for it appeared to occur on both sides of the base-line; however, it is thought that these were semi-permanent field-changes accompanied by radia-

[2] De Groot, Proc. Inst. Radio Eng., New York, 5, 1917, p. 75.

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tion-field. That they were not caused by rectification in the amplifier is shown by the fact that the deflections of this type were frequently small, and well within the linear amplification range. Further investigation of the semi-permanent field-changes was postponed until after the required number of radiation-fields had been examined; unfortunately, however, work in this direction was prohibited by the failure of the oscillograph tube. Regardless of sign, the mean semi-permanent field-change was 0.1372 volts per m. but this value is probably low, since off-scale deflections were common.

Though none of the semi-permanent field-changes measured were observed simultaneously with lightning flashes, owing to the impossibility of one observer's making both observations, yet it is known that the storms were in all cases over 50 km. distant. No thunder was heard, and the storms seemed to cling to the horizons, north, northwest, west, and southwest. Mostly they were just visible over the horizon, the lightning playing in the clouds. During the preliminary period, February to March 20, on one or two occasions two observers made observations; on these occasions it was found that every lightning flash did not produce a measurable field-change at the aerial, and conversely, every semi-permanent field-change was not invariably associated with visible lightning. Local storms are fairly high; one on December 31, 1926, at about 22^h was approximately 7 km. high as judged by the difference in time between the flash and the thunderclap.

Summary

Of nearly 1000 atmospheric radiation wave-forms observed over a period of 9 days at the Watheroo Magnetic Observatory from March 20 to March 28, 1927, the most frequently occurring form was found to be a negative peaked quasi-periodic consisting of 3 "half cycles," the maximum field change being associated with the second "half cycle." The mean duration and field-strength of this type, which occurred in 10 per cent of all cases examined, were 3458 μs. and 0.139 volts per m., respectively. Reconstructed drawings of the most frequently-occurring forms are given, as well as reproductions of actual sheets of observations.

The summer "roar" is shown to be most probably due to short-period, small-amplitude, almost continuous ripples, the period lying between 50 and 100 μs. and the amplitude being approximately 0.008 volts per m. The static "rattle" noted by

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De Groot in the Dutch East Indies is probably due to ripples similar to these.

Negative atmospherics predominated, the only notable positive being an aperiodic of relatively long duration, this form being almost invariably associated with lightning. Quasi-periodics were more numerous than aperiodics, and the peaked form occurred in greater numbers than the rounded. The symmetrical negative peaked aperiodic was noteworthy, because of the shortness of its duration, 786 μ sec. All atmospherics seen on the oscillograph screen were accompanied by a noise in a radio receiver.

Semi-permanent field-changes were fairly numerous, the negative change occurring far more frequently than the positive, the thunderstorms producing these changes being usually over 50 km. distant.

Acknowledgments

Thanks are due to Mr. H. F. Johnston, Observer-in-Charge at Watheroo for material assistance in the provision of batteries, of which there are many in the apparatus, and for continued interest in the work; to Mr. F. W. Wood who helped to identify the various time-base screen patterns with the aid of a violin; to the Department of Physics of the University of Western Australia for the loan of a tuning fork, and to Mr. D. W. Everson of that Department, for the manufacture of the sinusoidal oscillator.

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PROPAGATION OF SHORT WAVES DURING A SOLAR ECLIPSE

By

Edwin J. Alway

Special watch was kept on 30 to 45 meters from 0400 to 0530 hours G.M.T. on the morning of June 29, 1927, for the purpose of observing effect of solar eclipse on transmission of 40–45 meter waveband.

The receiving station was situated at Heliopolis near Cairo, Egypt. Dawn occurred at 0200 hours G.M.T. Progress of eclipse could be observed through window; eclipse was only partial in Egypt, though total in England at 0524 G.M.T. First indication of effect on short-wave transmissions was observed on signals from WIZ, 42 meters, which generally persists with gradually weakening strength until some hours after dawn. On this occasion, WIZ began to rise slightly in strength at 0420 hours, which was time of first contact on line of totality. At 0423 station ER5AA was heard calling, quite loudly, CQdeER5AA on 32 meters.

This station has never before been heard here as late as 0423. The wave in question, however, 32 meters, was not considered long enough to show any marked night-contra-daylight properties. Listening out was now confined to 40–45 meter band; signals from Europe on these waves always fade to zero when daylight covers whole path, daylight being 2 hours later in Europe.

At 0430, 10 minutes after first contact, a Vickers Vanguard aircraft GEBCP, was heard calling, "CQ de GEBCP—Time now 0530 B.S.T.—over Mersey—no bumps—over Mersey—" .... and later .... "—now over Liverpool—speech—std bi fone— ...."

This machine was transmitting on a wave of 42.5 meters, and was received R7; machine was then only a few miles from line of totality.

At 0450 the following was heard, "CQ de LCHO—test transmission from Oslo (Norway), radio on 45 meters," (R7).

At 0500 the British amateur G2WJ was heard on 45 meters sending code words. These code words have not yet been checked, but reception of them was not at all doubtful; here are two:

Original Manuscript received by the Institute, July 19, 1927.

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"UFEOFROGEYIMEOMAESYIZYOFAAROICU de G2WJ," and "YNIOCAIREUHAOZUYSAUMENAJAAGIUFY de G2WJ."

At 0500, WIZ was still heard quite loudly and persisted till after 0530 G.M.T.

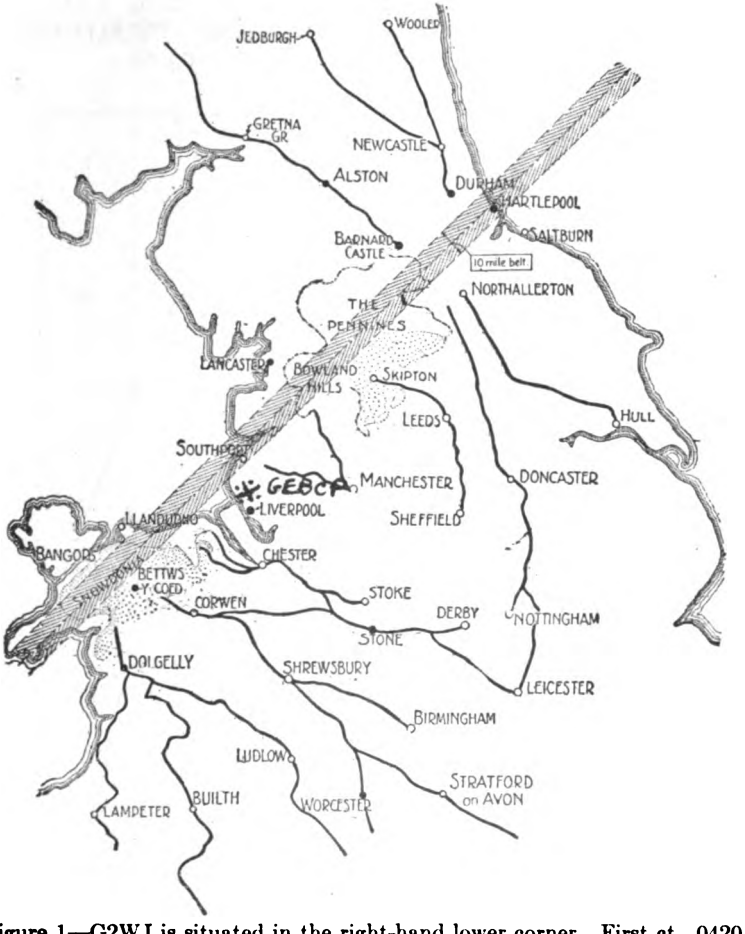

Figure 1—G2WJ is situated in the right-hand lower corner. First ct., 0420; total ct., 0524; last ct., 0614.

At 0516 G.M.T. the following call was heard on about 40 meters, "8 CER NU 3ASL—R ck—ur tone gud AC cut a little—gru hr—vy gld ti QSO ob, bi uw om so wl QRT nw vy best 73's cu agn—A 8 C.E.R. nu 3 AHL (?)." This transmission, apparently from an amateur station in the third zone in the United States,

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could scarcely have been heard at Heliopolis, Egypt, after three and one-quarter hours of daylight under normal circumstances. It is worthy of note that this particular station was heard just previous to the time of totality in England.

At 0520, G2WJ could still be heard sending test code words, but was not so easily receivable, more errors being made; the following incomplete code word was received: "TOALOORY-UZOAUUASMXA—de G2WJ."

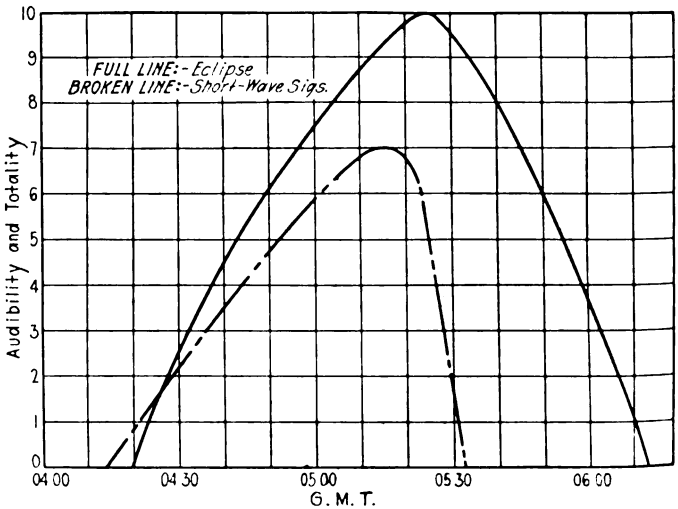

Figure 2

At 0530, Oslo LCIIO, had also fallen in strength, though still readable.

A station on 39 meters was also heard at 0530 calling "18GR de OC1," but I have been unable to identify the calls signs. After 0530 G.M.T., all signals ceased on 40–45 meters, and atmospherics, which had been present from 0400–0530 also ceased, and the watch was discontinued.

One concludes by assuming that, at least locally, a solar eclipse produces a pseudo-night-effect, this effect beginning with the eclipse but finishing before it; that is, when the eclipse occurs in the morning.

N.B. G2WJ is situated in London N 12.

GEBCP and G2WJ were approximately 2500 miles from Cairo.

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LCHO was approximately 3000 miles from Cairo.

ER5AA has not been identified, but is probably of Russian origin; if so, it is more than 500 miles distant.

8CER and 3AHL (or 3 ASL) were prefixed by the Amateur International Signs NU and are of American origin.

3 valve receiver used 1 detector, 2 audio-frequency amplifiers.

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Discussion on

LONG DISTANCE RECEIVING MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1925* (L. W. AUSTIN)

K. Sreenivasan†: It was hoped at one time that the observations on Madras radio should be examined with respect to magnetic disturbances; but the difficulty in getting magnetic data together with the assumption that at short distances like 300 Km, such disturbances will not have noticeable effect on reception prevented any further work.

The important paper by Mr. G. W. Pickard[1] on the correlation of radio reception with solar activity showed that the assumption was not justifiable. An examination of the intensity variations in relation to solar activity and terrestrial mangetism is now undertaken on the lines suggested by Mr. Pickard[2] in his comments on my discussion of Dr. Austin's paper.

The curve obtained from the Madras observations of October and November, 1926 show the marked influence of sunspot activity with reception. Considering that both these months were of comparatively violent changes in intensity from day to day and week to week, the inverse relationship as shown by the curve is of considerable interest. That the day observations of Bangalore are not an exception seems to be borne out by similar relationship that Mr. Pickard has found with Dr. Austin's observations on European stations at Washington.

Of peculiar interest is Mr. Pickard's remark that 1926 is an exception to the general rule of direct relationship between sunspots and day reception. Why this should be so, it is very difficult for me to understand. But 1926 was characterised by a violent magnetic storm during the latter part of October which completely blocked the beam system of transmission between England and Canada. According to published information available, there has been practically no abnormality during the severest part of the storm in the reception of long-wave signals. This indicates some kind of selective action. Without laying too much stress on the

Original Manuscript received by the Institute, October 5, 1927.

* PROCEEDINGS, Institute of Radio Engineers, **14**, 663, 1926.

[1] PROCEEDINGS, Institute of Radio Engineers, **15**, 83, 1927.

[2] PROCEEDINGS, Institute of Radio Engineers, **15**, 539, 1927.

† Radio Laboratory, Dept. of Electrical Technology, Indian Institute of Science, Bangalore, India.

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Madras observations, the weekly averages seem to suggest an explanation which may be worth some attention, although at present it can only be in the nature of a conjecture. Is it possible that short waves on account of their penetration to a greater height into the ionised atmosphere up above respond more readily than lower frequency waves to such phenomena as magnetic storms? Is it not likely that any violent disturbances above will reach the lower "turbid" layers rather slowly and as a consequence that long-wave reception will show less immediate and violent changes, beginning some time after the storm has started and assuming normal values comparatively long after it has ceased?

If there were sufficient data on reception over a number of years throughout the radio spectrum from short waves through the broadcast band to long waves, the above suggestion might have been examined for what it is worth.

Besides the Madras observations, the intensity of the British Post Office Station at Rugby (G B R, 16 Kc. per sec.) has been measured daily due to the kindness of Mr. E. H. Shaughnessy, who readily agreed to the transmission of special test signals twice daily for two months from July 10th, 1927. Each of them consists of a three-minute dash, the morning signal lasting from 0115 to 0118 G.M.T. and the afternoon signal from 1115 to 1115 G.M.T. With a time difference of about 5½ hours between Rugby and Bangalore, the path of the morning signal lies mostly in darkness. During the evening signal, the path is lighted throughout.

Report on these measurements will be made soon after the transmission stops after September 10th. Unfortunately, the measurements on the test signals from Malabar (P K X, 15.6 Km) at 2340 G.M.T. were discontinued from July 10th, 1927. Even as long as it lasted, there were so many breaks due to one reason or another, that the number of readings is far too small to be useful.

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THE RELATION OF RADIO RECEPTION TO SUNSPOT POSITION AND AREA*

By

GREENLEAF W. PICKARD

(Consulting Engineer, the Wireless Specialty Apparatus Company, Boston, Massachusetts.

(Communication from the International Union of Scientific Radio Telegraphy.)

As a matter of years or even months, the relation between radio reception and solar activity is now quite definitely known.[1] An increase of solar disturbance, as indicated by an increase in the number of sunspots, solar constant or ultraviolet radiation, is accompanied by an improvement in day reception and a lowering of night reception. By taking periodic daily averages for a number of solar rotations certain general day-to-day relations can be found; for example, the meridian transit of a sunspot group is usually accompanied by a depression of night reception and (with the exception of 1926 and the early part of 1927) a slight increase of day field.[2]

Night reception in that portion of the broadcast band between 750 and 1500 kilocycles and over east-west transmission paths shows the highest relation to measures of solar activity. But even with this most sensitive type of radio transmission, if intervals of less than a month are taken the degree of correlation rapidly decreases. Comparing night reception at Newton Centre, Massachusetts, from stations WBBM and WGN at Chicago, with Wolfer Provisional Sunspot Numbers over the interval January, 1926 to September, 1927, the following correlations are found:

Interval	Correlation	r/ε
Three months,	-0.92 ± 0.038	24.0
Two months,	-0.75 ± 0.093	8.1
One month,	-0.54 ± 0.10	5.4
Fifteen days,	-0.45 ± 0.084	5.4
Seven days,	-0.30 ± 0.06	5.0

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* Presented before the International Union of Scientific Radio Telegraphy, Washington, D. C., October 14, 1927.

[1] "The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism," PROCEEDINGS of the Institute of Radio Engineers, Vol. 15, No. 2, February, 1927.

[2] "The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism. II." *Bull. National Research Council*, No. 61, July, 1927. PROCEEDINGS of the Institute of Radio Engineers, Vol. 15, No. 9, September, 1927.

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When a day-to-day comparison of reception and Wolfer sunspot numbers is attempted, a very low correlation is found. If recognition is taken of the distance of the sunspots from the sun's center, by multiplying their area by $(1-\rho)$, where ρ is the distance from the center of the solar disk, a somewhat larger correlation coefficient results. A further improvement results when only spots within a few degrees of the central meridian are taken. But even when position as well as area is considered, the

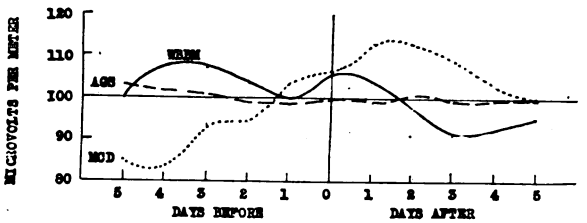

Figure 1—Reception and magnetism with maxima of 4 or more spots in central 13.3 deg. band. 88 cases, February 17, 1926 to May 21, 1927.

connection between individual sunspot groups and reception is not sufficient to show that we are dealing with cause and effect.

As a working hypothesis, I have assumed that there are definite solar areas which are responsible for reception disturbances, and which usually persist for a number of solar rotations. In general these active areas are also areas of sunspot formation, but they are effective in producing reception disturbances even when

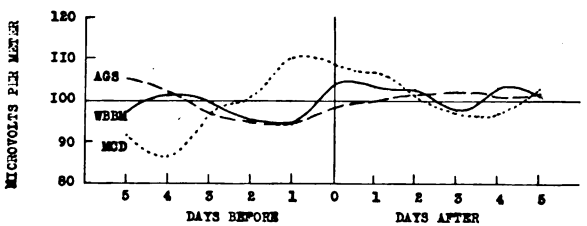

Figure 2—Reception and magnetism with maxima of 4 to 10 spots in central 13.3 deg. band. 62 cases, February 17, 1926 to May 21, 1927.

they contain no spots. Large sunspot groups may also occur entirely outside these active areas, in which event the earthward presentation of such spots is less likely to be accompanied by changes in reception.

But as the active areas persist for a number of solar rotations (sometimes over a year) and are the most likely places for sunspot formations, we may still use sunspot area and position to derive

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general relations despite their inadequacy in the individual case. It is the object of this paper to show that the central passage of sunspots does not exactly coincide with the associated reception disturbances, and that the interval between the transit of the sunspots and the reception change varies both with the area of the sunspot group and with the particular period taken for investigation.

In the first seven figures night reception, day reception, and terrestrial magnetism are compared with the central passage of sunspot maxima. The night reception is taken from my measurements at Newton Centre, Massachusetts, and Seabrook Beach, New Hampshire, of station WBBM at Chicago, the day reception

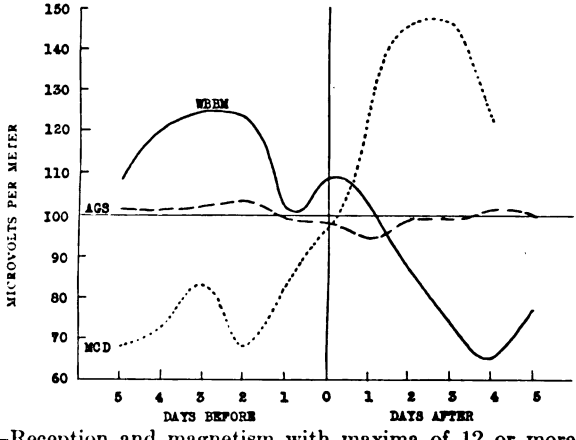

Figure 3—Reception and magnetism with maxima of 12 or more spots in central 13.3 deg. band. 26 cases May 13, 1926 to May 12, 1927.

is at Washington from AGS to Nauen, Germany, while the magnetic measure used is the Magnetic Character of Day Number given by the Observatory at Cheltenham, Maryland. The sunspots were counted in a central band 13.3 deg. wide, which is one day's travel in solar rotation.

In Fig. 1 is shown the relation between day and night reception, terrestrial magnetism and sunspot maxima of four or more spots, over the period February 17, 1927, to May 21, 1927. The ordinates represent percentage variation of the elements, and the graph covers an interval of eleven days. Night reception, shown in full-line, reaches a minimun about three days after the central passage of the spots, day reception an ill-defined minimum one day before, while magnetism, shown in dotted line, shows a maximum of disturbance between two and three days after.

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A comparison of reception and magnetism over the same period, but with sunspot maxima limited to four to ten spots is shown in Fig. 2. Now both night and day reception show minima one or two days before, with a maximum of magnetic disturbance one day before.

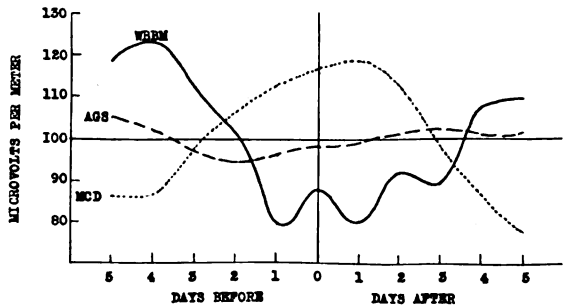

Figure 4—Reception and magnetism with maxima of 4 to 10 spots in central 13.3 deg. band. 26 cases, February 17, 1926 to October 4, 1926.

In Fig. 3 only the larger sunspot groups are taken, with spot numbers of twelve or more in the central band. The amplitude of the night reception and magnetic disturbance has now increased, with minimum and maximum values, respectively, four and three days after the passage of the spots, while day reception shows a slight depression one day after.

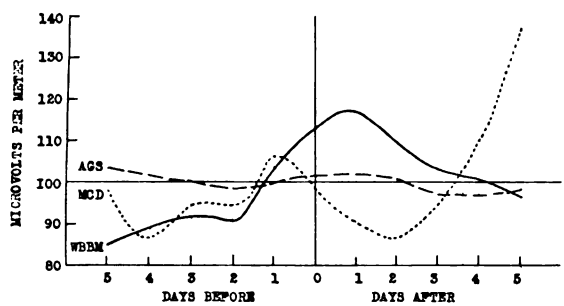

Figure 5—Reception and magnetism of 4 to 10 spots in central 13.3 deg. band. 36 cases, October 8, 1926 to May 21, 1927.

It would seem from these figures that the interval between sunspot transit and reception depression is related to sunspot area, much as if the smaller spot groups were on the western, and the larger spots on the eastern boundary of the active areas. But there is another factor to be considered, namely the period covered by the observations. In the preceding figures this has been approxi-

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mately fifteen months; now this interval will be broken into two nearly equal parts. In Fig. 4 the interval is from February 17, 1926 to October 4, 1926, with sunspot maxima of four to ten. Night reception shows a depression centering one day after, magnetism a maximum one day after, and day reception a minimum two days before.

In Fig. 5 the period is October 8, 1926 to May 21, 1927, with sunspot maxima of four to ten. Although at first sight this figure is the inverse of the preceding one, the night reception maximum one day after the passage of the spots is not related to this particu-

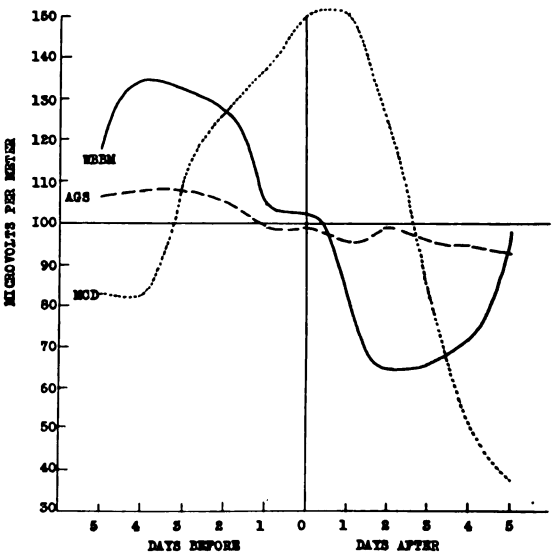

Figure 6—Reception and magnetism with maxima of 12 or more spots in central 13.3 deg. band. 8 cases, May 13, 1926 to September 19, 1926.

lar transit, but merely represents reception recovery from some preceding disturbance; actually what has happened is merely a considerable displacement in time. The inverse relation of magnetism and night reception is here the best guide to the true position of the reception minimum, which is more than five days after the spot transit. Apparently the smaller sunspot groups were quite near the center of the active areas during the first seven months, but in the second period they moved east by at least 65 deg.

The larger sunspot groups show a similar displacement, although to a lesser degree. In Fig. 6 reception and magnetism from May 13, 1926 to September 19, 1926, are compared with sunspot

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maxima of twelve or more. Night reception shows a minimum two days after, and magnetism a maximum one day after. Fig. 7 is also for sunspot maxima of twelve or more, for the period of October 11, 1926 to May 12, 1927, and here the reception and magnetic elements show minimum and maximum values three and four days, respectively, after the transit of the spots; a displacement of about two days or 27 deg. to the east.

It is evident from the preceding figures that the relation of reception to magnetism is much more intimate than its connection

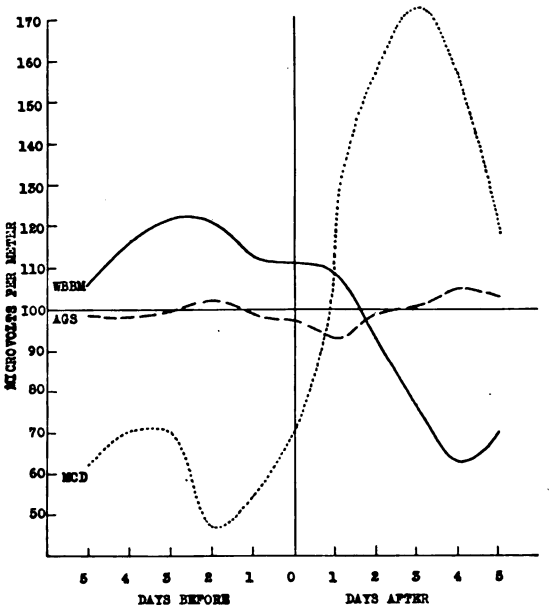

Figure 7—Reception and magnetism with maxima of 12 or more spots in central 13.3 deg. band. 18 cases, October 11, 1926 to May 12, 1927.

with sunspots. In Fig. 8 the magnetic storm is taken as the event, instead of sunspot transit. Within the nineteen months from January, 1926 to August, 1927, there were fourteen magnetic storms recorded by Cheltenham, and the depression of night reception accompanying these storms is very striking. Day reception shows an inverse effect, an increase of field accompanying and following the storm. The night reception maximum three days before is not in anticipation of the coming storm; it merely represents the recovery of reception from the depression associated with a preceding magnetic disturbance.

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Although the most marked reception disturbances accompany magnetic storms, even a moderately disturbed day has its associated reception depression. In Fig. 9, which covers the period January 24 to June 12, 1927, all magnetically-disturbed days are taken, that is, all days with character number of 1 or 2. The night reception data are here an average of three transmission paths; WBBM at Washington and Newton Centre, and WJAX of Jack-

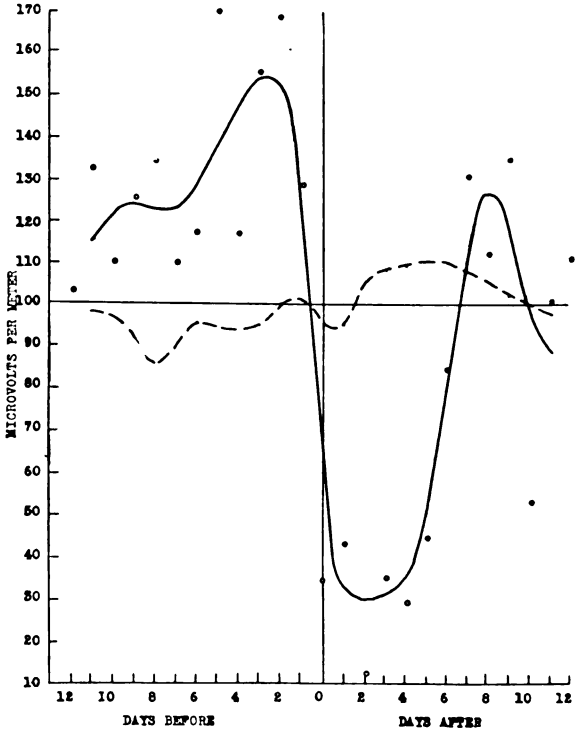

Figure 8—Reception from WBBM and WGN with magnetic storms observed at Cheltenham. 14 cases, January 26, 1926 to August 19, 1927. AGS day-reception in broken line.

sonville, Florida, at Washington. Over the first half of 1927 magnetic disturbances came at intervals of approximately fifteen days, so the reception maximum three days before represents the recovery of reception from the depression associated with a preceding magnetic disturbance, and the curve has a period of fifteen days.

On the 19th and 20th of August, 1927, there was a magnetic storm of considerable intensity, with displays of Northern Lights on both evenings visible throughout New England. In Fig. 10 I

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have shown the reception depression associated with this storm, in which the field from WGN remained at a low value for a week.

Night reception over any extended period is largely a series of abrupt depressions followed by relatively slow recoveries. The recovery, at least during a sunspot maximum, is rarely completed before the next depression occurs, so that the mean field for, say,

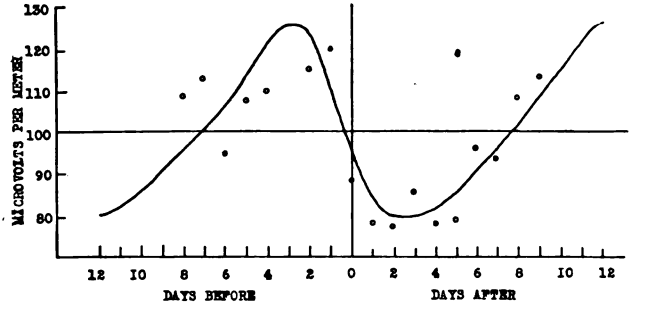

Figure 9—Mean reception from WBBM and WJAX at Washington and Newton Centre with maxima of magnetic disturbance at Cheltenham. 36 cases, January 24 to June 12, 1927.

a month, is largely determined by the interval between these abrupt depressions. If this interval is short, for example a week, the mean field will be low; if two or three quiescent weeks intervene the mean field will be high.

These depressions quite definitely accompany disturbances of terrestrial magnetism, and less definitely the central passage of

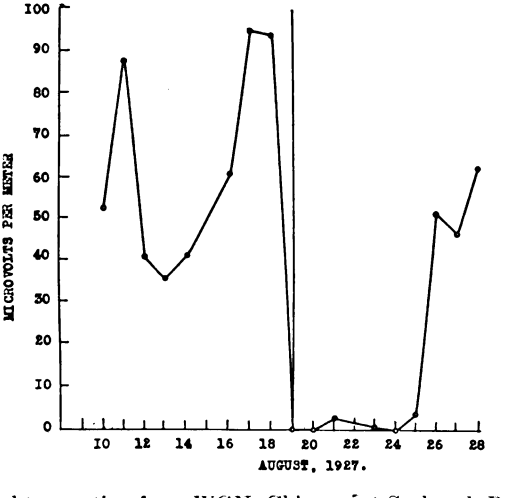

Figure 10—Night reception from WGN, Chicago, at Seabrook Beach, N. H., and the magnetic storm of August 19 and 20, 1927.

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sunspot groups, but in neither case are we warranted in assuming that the connection is one of cause and effect. However, the relation of reception to terrestrial magnetism is so definite it is difficult to escape the conclusion that disturbances of these two elements arise from a common and nearby cause; an electrical change in our atmosphere. Any inexactness in this relation might be explained on the assumption that magnetic disturbances require an electrical movement, whereas reception is dependent upon the degree and disposition of ionisation.

I wish to acknowledge the cooperation of Mr. H. H. Clayton, who not only made a preliminary analysis of my reception data by the method which I have used herein, but also supplied from his own solar observations the central band sunspot numbers. I am also indebted to the U. S. Coast and Geodetic Survey for the Magnetic Character of Day Numbers from the Cheltenham Observatory.

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ABBREVIATED METHOD FOR CALCULATING THE INDUCTANCE OF IRREGULAR PLANE POLYGONS OF ROUND WIRE

(Part I of paper, "On the Calculation of Closed Aerials")

By

V. J. Bashenoff

(Chief of the Wireless Section of the State Electrical Research Institute, Moscow, Union of Socialist Soviet Republics)

Introduction

There are many well-known theoretical and experimental papers dealing with the calculation of the inductance of coils and loops of small area and large number of turns. Such coils, arranged so as to allow for their rotation about a vertical axis, have been used for portable antennas and are suitable for installation in the interior of a building. The problem of the calculation of the constants of closed loops having a large area and a very small number of turns has, on the contrary, received until the present very little attention in radio journals or radio handbooks.

Such antennas, which often consist of a single turn of wire supported by a single mast, are of especial importance at present on account of the definite tendency, very evident in western Europe, at least, to adopt this type of antenna in preference to the smaller loops which have earlier been almost exclusively used. For example, there may be cited changes made in 1924 in the radio centrals of Paris and Berlin (Creisne and Geltow). These stations, which in the period 1920–23 employed small loops with an area of a turn of from 4 to 16 square meters, now make use of closed aerials with an area of several thousand square meters, hung from masts as high as 75 meters. The Marconi Wireless Telegraph and Telephone Company employs in all its radio receiving sets a radiogoniometer system consisting of two mutually-perpendicular triangular closed aerials of one or two turns, hung from masts which, in different cases, range from 12 to 75 meters in height.

Finally, the French Society, "S.F.R." in its 1926 model of universal direction-finder for wavelengths of from 100 to 3000 meters has come over to the use of a large loop of only four turns,

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in place of the small loops of a large number of turns which it used formerly.[1]

The superiority, both from practical and scientific standpoints, of closed aerials of a few turns of great area, over loops of many turns of small area was established, as far back as 1915, by the author of the present paper, as a result of an experimental and theoretical study of radio reception with (a) coils with areas of a single turn up to 20 square meters, and (b) closed aerials with areas of some thousands of square meters, supported by a single mast.

This principle[2] has been embodied in the design of some dozens of radio stations installed since that time under the direction of the author. As an example may be mentioned the construction in 1920–4 of the radio station of Lubertzy[3] which was the main receiving station of the Radio Central of Moscow.

In all the radio stations built by the author the closed aerial is hung from a single mast, a construction dictated by economical and practical considerations. As to the shape of such aerials a triangular form would give the greatest area, but in practice to reduce the effective resistance and capacity it is necessary to cut off the lower corners of the triangle.[4]

For purposes of approximate calculation we may seek to replace the irregular polygons, formed by such closed aerials, by a simpler equivalent plane figure of either equal area or equal perimeter. Of such figures it is obvious that a triangle most closely approximates the given contour. Exceptions would be offered by closed

[1] This coincides with the conclusion expressed by the Radio Research Board, Department of Scientific and Industrial Research, U. K., Special Report, No. 1, 1923, that a single-turn loop is the theoretical ideal for a direction finder.

[2] The comparison of the reception on closed and frame aerials will be treated in a later paper. On this account a simple note will suffice for the present to show the theoretical advantage of the closed aerial of large area and few turns over the coil of small area and many turns. The product Sn which occurs in the expression for the effective height of a frame aerial may be obtained for a given length l of wire, either with a single turn aerial or by using a frame of n turns. Let us suppose, for simplicity, that in both cases enclose a square area. Then for a closed aerial of a single turn the length of a turn is $\frac{l}{4}$ and the area $s=\frac{l^2}{16}$. For the frame antenna of n turns the length of a side is $\frac{l}{4n}$ and the area $s_n=n$ times the area of a single turn, or $\frac{l^2}{16n}$. Thus the effective height is diminished in proportion as the number of turns is increased.

[3] German Patents, No. 420450, Oct. 26, 1925 and No. 430695, June 18, 1926.

[4] Telefunken Patent D. R. P. No. 317880, and the author's patent D. R. P. No. 420450.

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aerials suspended from a single point and having circular, rectangular or square shapes. The first two cases are not met in practice, and the last, patented by the Telefunken Co. in a small frame aerial, offers no advantages over the truncated triangular shape.

Because, then, of the predominant importance of the triangular shape, the author and his collaborators have, as will be seen later, investigated especially the problem of the calculation of the constants of triangular antennas. The following is an attempt to indicate methods for the calculation of the inductance of closed aerials. The accuracy of the theoretical analysis is confirmed by experimental data obtained by the author over a period of ten years.

Calculation of the Inductance—Single Turn Closed Aerial

A. Special Formulas for Various Figures

The following formulas for several of the geometrical figures are well-known. All the formulas are based on: (1) Circular section of wire with radius r. (2) Wire-material having $\mu = 1$—unless some other value is specified. (3) Naperian logarithms.

1. Circle

The formula giving the greatest accuracy (for direct and low-frequency currents) is, with a—radius of the circle,

$$L = 4\pi a\left[\left(1+\frac{r^2}{8a^2}\right)\log\frac{8a}{r}+\frac{r^2}{24a^2}-1.75\right] \qquad (1)$$

(Rayleigh and Niven)

Approximate formulas, generally used in practice: (a) for low frequencies

$$L = 4\pi a\left(\log\frac{8a}{r}-1.75\right) \qquad (2)$$

(Kirchoff)

(b) for high frequencies

$$L = 4\pi a\left(\log\frac{8a}{r}-2\right) \qquad (3)$$

(c) general formula—for any frequency

$$L = 4\pi a\left(\log\frac{8a}{r}-2+\mu\delta\right) \qquad (4)$$

(Bureau of Standards)

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in which δ—one of the correction factors for frequency is determined as a function of x (see Fig. 1)—which in turn is expressed by

$$x = 0.281 \cdot r\sqrt{\frac{\mu \cdot f}{\rho}} \tag{5}$$

In this formula: a—radius of circle; f—frequency; r—radius of wire in centimeters; μ—magnetic permeability ($\mu = 1$ for all

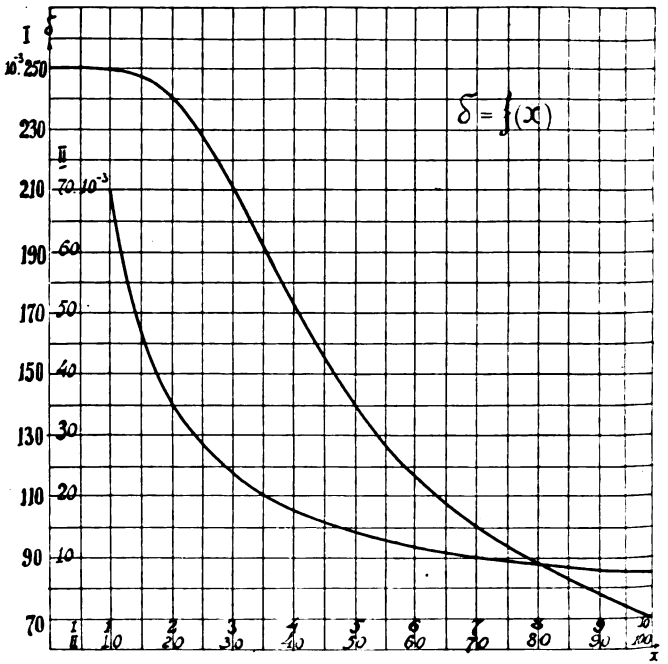

Figure 1

wires except ferromagnetics); ρ—specific resistance of wire in microohms/cm³.

2. Square

$$L = 8a\left(\log\frac{a}{r} + \frac{r}{a} - 0.774 + \mu\delta\right) \tag{6}$$

(Kirchoff)

in which a is the side of the square.

Another approximate formula is

$$L = a.\gamma \tag{7}$$

(Esau)

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in which γ is a function of $\frac{a}{r}$ and is obtained from Fig. 2.

3. *Equilateral Triangle*

The formula is:

$$L = 6a\left(\log\frac{a}{r} - 1.55465 + \frac{r}{a} 1/4\frac{r^2}{a^2} + \cdots\right) \tag{8}$$

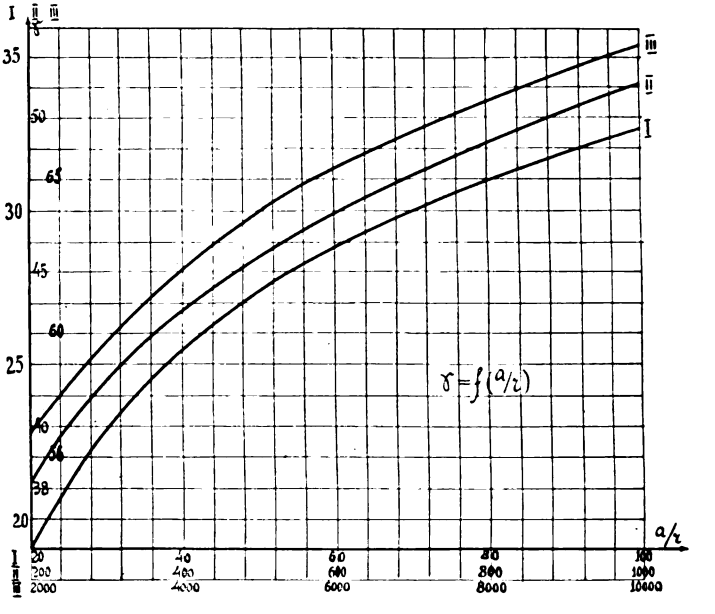

Figure 2

for direct current (low frequencies) (Grover).[5]

$$L = 6a\left(\log\frac{a}{r} - 1.405465 + \mu\delta\right)\cdots \tag{8a}$$

for radio frequencies.

4. *Regular Hexagon*

The formula is

$$L = 12a\left(\log\frac{a}{r} + 0.098476 + \frac{r}{a} - 1/4\ \frac{r^2}{a^2} + \cdots\right) \tag{9}$$

[5] The formulas (8), (9), and (10) were found by Grover in deriving a formula for calculating the inductance of polygonal coils, *Sci. Paper 468*, Bureau of Standards.

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for direct current (low frequencies) (Grover)[5]

$$L=12a\left(\log\frac{a}{r}-0.151524+\mu\delta\right)\cdots \tag{9a}$$

for radio frequencies.

5. Regular Octagon

The formula is

$$L=16a\left(\log\frac{a}{r}+0.461976+\frac{r}{a}-1/4\ \frac{r^2}{a^2}+\cdots\right) \tag{10}$$

for direct current (low frequencies) (Grover)[5]

$$L=16a\left(\log\frac{a}{r}+0.211976+\mu\delta\right) \tag{10a}$$

for radio frequencies.

6. Rectangle

$$L=4\left[(a+b)\log\frac{2ab}{r}-a\log(a+d)-b.\log(b+d)\right]$$
$$+4[\mu\delta(a+b)+2(d+r)-2(a+b)] \tag{11}$$

in which a and b are the sides of the rectangle and d its diagonal.

7. Triangle

(a) *Right triangle.*

The following formula gives a degree of accuracy amply sufficient for all practical purposes.

$$L=a\left[\frac{a}{c}\log\frac{a(c-a)}{b(c+b)}+\log\frac{4a^2b^2}{r^2(c+a)^2}\right]+b\left[\frac{b}{c}\log\frac{b(c-b)}{a(c+a)}\right.$$
$$\left.+\log\frac{4a^2b^2}{r^2(c+b)^2}\right]+c\left[\log\frac{a^2b^2c^2}{r^2(a^2+c^2)(b^2+c^2)}-2\right] \tag{12}$$

(Bashenoff)[6]

(b) *Equal-leg right triangle.*

This formula is easily derived (see below)

$$L=2l\left[\log\frac{2l}{r}-3.331\right] \tag{13}$$

in which l is the perimeter of the triangle.

[6] For the derivation of formula (12) see *Jahrb. draht. Tel.* 1926, vol. 27, no. 4 and "Transactions of the State Electrical Research Institute" (Moscow) publ. 14.

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8. Quadrilateral (trapezium)

Of definite shape (Fig. 3).

$$L=2l\left[\log\frac{2l}{r}-3.227+\mu\delta\right] \qquad (14)$$

(Starik).[7]

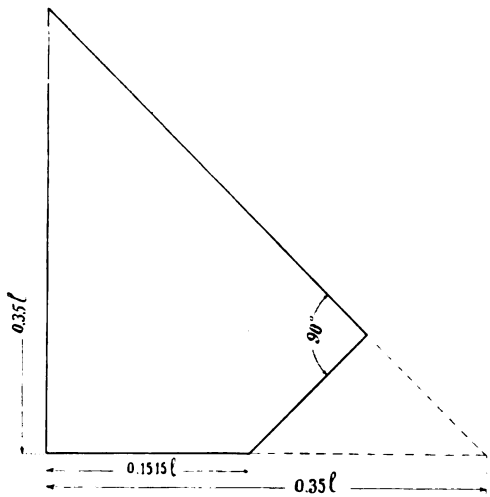

Figure 3

B. General Formula for a Plane Polygonal Figure Having No Reentrant Angles, Including Oblique-Angled Triangles

1st Method

The approximate method of computing the inductance of such polygonal figures is based on the author's assumption,[6] that the inductance of an oblique-angled triangle or another plane figure must, practically, be very nearly equal to the inductance of a right triangle, having perimeter and area equal to those of the given plane figure.

Indeed, in selecting an equivalent simple figure whose inductance shall approximate that of the actual (more complicated) figure, the most important condition to be satisfied is that of *equal perimeter*. This insures that the sum of the *self* inductances of the various wires of which the figure is composed shall be nearly the same both in the actual and the equivalent figures.

[7] Formula (14) was derived by Mr. Starik, by summation of the self and mutual inductances of all the elements, using the well-known formula given below.

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The total inductance of the figure is less than the sum of the self inductances of the separate elements, being reduced by the mutual inductances between certain of the elements of the figure. The mutual inductance of any pair of elements is greater the nearer the wires are together, and the more closely parallel they are. Thus, the mutual effects will be approximately taken into account by imposing the further condition that the area of the actual and the equivalent figures shall be the same. Since, however, in most practical cases the sum of the mutual inductances amount to considerably less than half the value of the sum of the self inductances of the wires, this second condition is not so

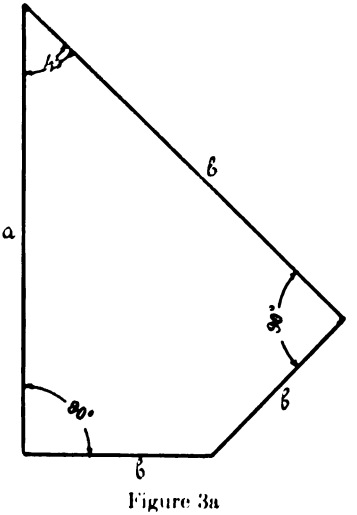

Figure 3a

important as that of equal perimeter. In those cases where both conditions cannot be simultaneously satisfied, that of equal perimeter should be chosen, and the area of the equivalent figure should be taken as nearly equal to that of the actual figure as possible. This amounts to imposing the condition that the right triangle of equal *perimeter* shall have equal legs.

With this assumption formula (12) may be used, substituting for a, b, and c the values found in replacing the given figure by a right triangle of equal perimeter and area; the sides a and b of the right angle of this equivalent right triangle are expressed by

$$a = \frac{8S}{\left(l+\frac{4S}{l}\right) + \sqrt{\left(l+\frac{4S}{l}\right)^2 - 32S}} \qquad (15)$$

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and

$$b=\frac{1}{4}\left[\left(l+\frac{4S}{l}\right)\pm\sqrt{\left(l+\frac{4S}{l}\right)^2-32S}\right] \tag{16}$$

in which S is the area and l the perimeter of the figure.

This replacement is possible only for figures for which there exists between l and S the following relation

$$\frac{l}{\sqrt{S}}\geq 4.8284 \tag{17}$$

If the perimeter l and the area S of the figure satisfy the expression

$$\frac{l}{\sqrt{S}}<4.8284 \tag{18}$$

it is necessary to resort to a roughly approximate method of calculation, using formula (14), in which a is one of the legs of the equal-leg right triangle which has a perimeter equal to that of the given plane figure.

Such a triangle always exists and its sides are determined by the following formulas:

$$a=b=\frac{l}{2+\sqrt{2}};\quad c=\frac{l}{1+\sqrt{2}} \tag{19}$$

This approximation method (formulas (18) and (19)) gives generally a lower value than the measured value by 5 to 20 per cent. In general, the more the ratio $l/\sqrt{s}$ for a given figure falls short of 4.8284 the more the calculated value of l exceeds its true value.

Dr. Grover has suggested to me the possibility of extending the range of values of the ratio $l/\sqrt{s}$ for which an equivalent figure may be found. If, instead of a triangle, a rectangle be sought which shall be equivalent to the given figure, it is found that the conditions of equal area and perimeter may be satisfied, provided that the ratio $l/\sqrt{s}$ is not less than 4. Thus for values of $l/\sqrt{s}$ between 4 and 4.8284 more accurate values of the inductance are obtained if a rectangle of equal perimeter and equal area be used as the equivalent figure, than when an equal-leg right-angled triangle of equal perimeter is employed. For values of $l/\sqrt{s}\geqq 4.8284$ the use of the equivalent right triangle is to be recommended.

As an illustration of these points Dr. Grover has communicated to me a calculation of the inductance of one of the antenna forms described in one of my former papers,[7] viz., one of the Lübertzy

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closed aerials, shown in Fig. 4, for which the perimeter is 370.9 meters and the area is 6658.5 square meters.

The accurate value of the inductance was obtained by summation of the self inductances of the different elements (using the well-known formula for the self inductance of a straight wire) and the mutual inductances of the various pairs of filaments. These mutual inductances can be calculated by the general formulas of Martens[8] or Campbell.[9] The calculations were not difficult but rather long and tedious on account of the number of mutual inductances to be calculated. The value found (not including internal linkages) is 720140.

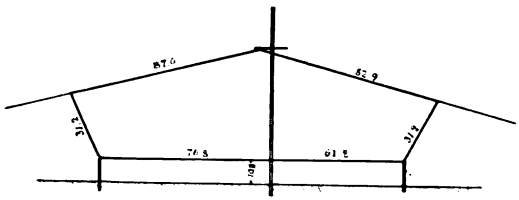

Figure 4

The following table shows the values obtained from the different approximations.

TABLE I

	l	$\frac{l}{\sqrt{s}}$	a	b	c
True Value	720140	4.546			
Equivalent Rectangle	722040	4.546	136.76	48.68	
Equilateral Triangle	714050	4.559	123.63		
Equal-Leg Right Triangle	704200	4.828	108.64	108.64	153.6
Regular Hexagon	755455	3.722	61.83		

The rectangle is the only figure for which both the perimeter and the area can be made equal to those of the given figure, although the equilateral triangle of the same perimeter has nearly the correct area in this case. The area of the equal-leg right triangle of the given perimeter is considerably smaller than that of the given figure, and a too small inductance is to be expected. The regular hexagon of equal perimeter has an area considerably too large, and thus an inductance too large is to be expected. We see that these differences are in the opposite direction to the values of the ratio $l/\sqrt{s}$. For this example the equivalent rectangle gives an approximation which is amply sufficient.

[8] Martens *Ann. der Phys.*, **29**, 963; 1909.
[9] Campbell, *Phys. Rev.*, June 1915.

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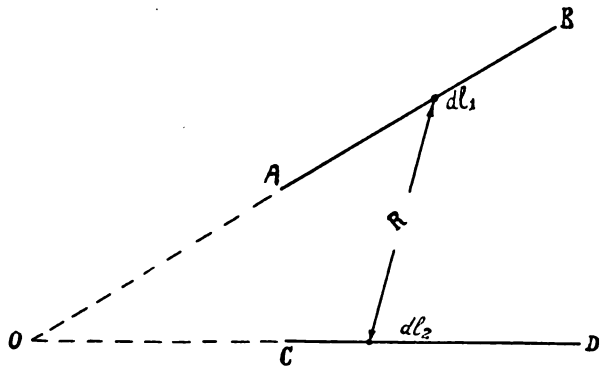

Figure 5

2nd Method

The following formula is taken as a basis for solving the problem

$$L = \sum_{i=1}^{i=n} L_i + \sum_{i=1}^{k=n} \sum_{k=1}^{k=n} M_{ik} \tag{20}$$

in which L_i is the coefficient of self-induction for one of the sides of the polygon; M_{ik}—coefficient of mutual induction between sides (i, k) of the same polygon.

As is well known, for a straight wire (using the same symbols)

$$L = 2l\left[\log\frac{2l}{r} - 1 + \mu\delta\right]$$

(Neumann)

As to M_{ik} for parallel wires, the distance between which is d

$$M = 2\left[l.\log\frac{l+\sqrt{l^2+d^2}}{d} - \sqrt{l^2+d^2} + d\right] \tag{22}$$

If l is large as compared with d

$$M = 2l\left[\log\frac{2l}{d} + \frac{d}{l} - 1\right] \tag{23}$$

For wires l_1 and l_2, intersecting at an angle e

$$M = \int_{l_1}\int_{l_2}\frac{dl_1 dl_2}{R}\cos e \tag{24}$$

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in which formula dl_1 and dl_2 are elements of the wires l_1 and l_2, and R—the distance between dl_1 and dl_2.

Lastly, if the wires do not intersect (see Fig. 5)

$$M_{\substack{AB\\CD}} = M_{\substack{OB\\OD}} - M_{\substack{OD\\OA}} + M_{\substack{OC\\OA}} - M_{\substack{OB\\OC}} \tag{25}$$

Integrating (24),

$$M = -\cos e \left\{ (a+b) \operatorname{arc\,sin} h\beta + a.\ \operatorname{arc\,sin} h \frac{C - a\cos e}{a \sin e} + b \operatorname{arc\,sin} h \frac{a - b\cos e}{b \sin e} \right\} \tag{26}$$

(Starik)

or, substituting $\qquad m = \frac{b}{a}; \quad \beta = \operatorname{cotan} e.$

$$M = -a\cos e \left\{ (1+m) \operatorname{arc\,sin} R\beta + \operatorname{arc\,sin} h \frac{m - \cos e}{\sin e} + m \operatorname{arc\,sin} h \frac{1 - m\cos e}{m \sin e} \right\} = -a \cdot \cos e \cdot f(m, e) \tag{27}$$

(Starik).[10]

It follows that M is equal to the length of one of the sides multiplied by a certain function of the ratio of the sides and the angle between them. By plotting this function the computation of the coefficient of mutual induction of each pair of intersecting wires is greatly simplified.

The calculation of the inductance of a plane polygonal figure may thus be made by formula (20) by calculating the self-inductance of the straight wires by (21) and the mutual inductances between them by formulas (22–27).

The calculation of L for a polygonal figure by the above method is rather long, as it requires a number of preliminary calculations. The author and one of his pupils, Mr. Starik, propose the following practical approximate method of computing the inductance.

The first term in the right side of the expression (20) takes into account the self-inductances of the sides of the polygon. We suppose the sum of these terms to be a function of l and l/r only and that they can be expressed as the self-inductance of a straight wire having the same radius of section and a length equal to the

[10] For the same solution, but in another form, see Mesny: "L'Usage des Cadres et la Radiogoniometrie."

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perimeter of the figure. Then, the first term in the right side of the expression (20) can be written[11] as:

$$\Sigma L_i = 2l\left(\log\frac{2l}{r} - 1 + \mu\delta\right) \tag{28}$$

in which, as before, $\mu\delta$ is a correction for frequency. It follows that for a given frequency and wire material

$$\Sigma L_i = f\left(\frac{l}{r}\right) \tag{29}$$

The second term in (20), expressed in its most general form (27) is

$$\Sigma M_{ik} = \Sigma a_{ik} f(e_{ik},\ \ m_{ik}) = 2ly \tag{30}$$

in which y for a given shape (i.e., for all similar figures having the same ϵ_{ik} *and* $M_{ik} = ai/ak$*) is a constant.*

For any plane polygonal figure in general we have, accordingly,

$$L = 2l\left(\log\frac{2l}{r} - 1 + \mu\delta + y\right) = 2l\left(\log\frac{2l}{r} - ak + \mu\delta\right) \tag{31}$$

in which a_k for a figure of given shape (definite ϵ_k and $m_{ik} = ai/ak$) is a constant.

Between two figures of equal perimeter and equal radius of wire according to (31) there can exist, at the same frequency f, only a difference in the parameter a_k. The value of the parameter a_k in its turn, as follows from the proposition set down above, must be the same for all figures with equal $l/\sqrt{s}$ when their

[11] The formula (28) is, of course, approximate, since it does not represent accurately the sum of the inductances of the sides. For example, for a triangle of sides a, b, c, the expression for the sum of the self inductances of the sides:

$$L_a + L_b + L_c = 2a\left(\log\frac{2a}{r} - 1 + \mu\delta\right) + 2b\left(\log\frac{2b}{r} - 1 + \mu\delta\right) + 2c\left(\log\frac{2c}{r} - 1 + \mu\delta\right) \tag{28a}$$

is not the same as $2l\left(\log\frac{2l}{r} - 1 + \mu\delta\right)$, where $l = a + b + c$. The quantity $2l\left(\log\frac{2l}{r} - 1 + \mu\delta\right)$ is the self inductance of a straight round wire of length l and this is accurately equal to

$$L_a + L_b + L_c + 2M_{a,b} + 2M_{a,c} + 2M_{b,c} \tag{28b}$$

in which the elements a, b, c all lie in the same straight line.

However, the equation (31) below is of the correct *form* to represent the inductance of any plane polygonal figure, if only the value of a_k can be properly (experimentally) determined.

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perimeters are equal, i.e., the parameter a_k is a function of this ratio only.

C. Verification of the Principles of Section B

Let us verify the validity of the principles stated in Section B by applying them to the special formulas given in section A. It must be noted to begin with that the assumptions made above limit the accuracy obtainable in applying the general formula (31); the aim of the latter is, however, to obtain approximate values having a sufficient degree of accuracy in practice for calculating closed aerials.

In closed aerials r/a is always small as compared to $\log a/r$; r (radius of the wire) very rarely exceeds 0.25 cm.; a (radius of the circle, side of the regular polygon, leg of equal-leg right triangle etc.) is, on the contrary, never less than 100 cm. The value of r/a is consequently always less than 0.0025, and may be certainly regarded as negligible in comparison with $\log a/r$ (which, under limitations just specified, is always >5.99). The corresponding difference of value does not exceed 0.4 per cent. We may certainly with still more right neglect the second and higher powers of the ratio r/a.

The special formulas of Section A, omitting the terms of the order of r/a, can be written as follows:

1. *Circle*

$$L=2l\left\{\log\frac{2l}{r}-2.451+\mu\delta\right\} \tag{32}$$

2. *Square*

$$L=2l\left\{\log\frac{2l}{r}-2.853+\mu\delta\right\} \tag{33}$$

3. *Equilateral triangle*

$$L=2l\left\{\log\frac{2l}{r}-3.197+\mu\delta\right\} \tag{34}$$

4. *Regular hexagon*

$$L=2l\left\{\log\frac{2l}{r}-2.636+\mu\delta\right\} \tag{35}$$

5. *Regular octagon*

$$L=2l\left\{\log\frac{2l}{r}-2.5610+\mu\delta\right\} \tag{36}$$

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6. *Polygon of the type shown in Fig. 3*

$$L=2l\left\{\log\frac{2l}{r}-3.227+\mu\delta\right\} \tag{14}$$

7. *Equal-leg right triangle*

$$L=2l\left\{\log\frac{2l}{r}-3.332+\mu\delta\right\} \tag{37}$$

8. *Regular pentagon (a new formula derived by Dr. Grover) with* $l/\sqrt{s}=3.812$

$$L=2l\left(\log\frac{2l}{Z}-2.712+\mu\delta\right) \tag{37a}$$

9. *Quadrilateral of the type shown in Fig. 3a* ($a=(\sqrt{2}/4)$, $l=0.35355l$; $b=(2-\sqrt{2})/4$, $l=0.14645l$) *(derived by Dr. Grover)*

$$L=2l\left(\log\frac{2l}{r}-3.091+\mu\delta\right) \tag{37b}$$

Using formulas (32–37), we can draw the curve $a_k=f\left(\frac{l}{\sqrt{s}}\right)$ in which $a_k=1-y$, by plotting the points for the above nine figures, for which the value of the ratio $\frac{l}{\sqrt{s}}$ is well-known.

The curve Fig. 6 was obtained in this manner. Further details of its construction are given below in Section D. From the curve, a_k can be read with sufficient approximation for a given ratio $l/\sqrt{s}$ and by using the value thus found in formula (31) the inductance can be calculated.

As to the quantity $\mu\delta$ found by means of a curve (Fig. I) in which it is given as a function of x, the expression (5) in the case, most often met with, of a copper wire at 20 deg. C, goes over into

$$x=0.2142r\sqrt{f} \tag{38}$$

This expression can be plotted in a nomogram (an abac) which makes practical calculations easier.

In practical calculations of closed aerials, constructed nearly always of non-magnetic metals $\mu=1$. The wire being of comparatively large section ($d\geqq 0.2$ cm.), $x=0.2142r\sqrt{f}$ (for copper at 20 deg. C) gives numerical values of 23 and more, even when $\lambda=25000$ m ($f=12000$). Consequently δ and $\mu\delta$ (at $\mu=1$) in Fig. 1 have values of about 0.03 or less. In closed aerials $\log 2l/r$

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has nearly always a larger value than 6 so that it follows that we can neglect $\mu\delta$ in wireless practice with a degree of accuracy never below a fraction of one per cent, provided we are interested in knowing the inductance for radio frequencies only. Thus the formula for the inductance of any plane figure at radio-frequencies is finally expressed by

$$L \cong 2l\left(\log\frac{2l}{r} - a_k\right) \tag{39}$$

in which l is the perimeter and a_k is a function of only $l/\sqrt{s}$ and is taken from the curve in Fig. 6.

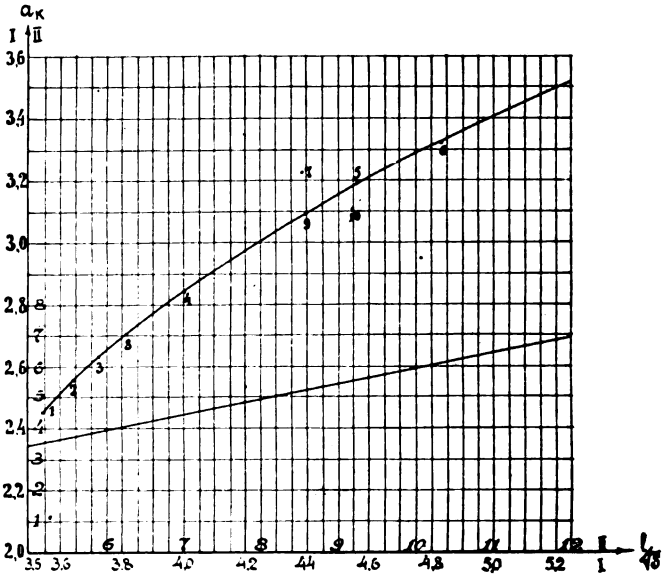

Figure 6

It is clear from formula (31), that the total change in the value of the inductance of a figure depends upon the value of ratio-perimeter to radius of cross section and value of $\mu\delta$; e.g., for a wire of a length 50 times that of its radius, the inductance is diminished only 6 per cent when the frequency changes from zero to infinity; for a wire with the ratio $l/r = 200000$ when the frequency change in the same limits, ΔL, is only 2 per cent.

The above formulas cover the calculation of all practical types of single-turn closed aerials. It should be noted, however, that figures having sharply reëntrant angles cannot be so calculated. This can be seen from Table II, which gives the measured values

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of the inductance found for certain irregular figures having reëntrant angles, together with values of the inductance calculated by formula (35) and curve 6.

TABLE II

Figure (see Fig. 7)	Perimeter of figure cm	Area of figure cm²	$\frac{l}{\sqrt{s}}$	L measured	L calc. from (35) and curve (6)	ΔL in per cent
1	2314	186000	5.43	43000	35090	−18.04
2	4000	454500	5.93	62500	60975	− 2.44
3	4106	304500	7.54	59000	56925	− 3.52
4	4332	154500	11.03	37500	43945	−23.6
5	4332	454500	6.42	64500	64650	+ 0.23

NOTE: For all figures $r = 0.1$ cm, $\mu = 1$.

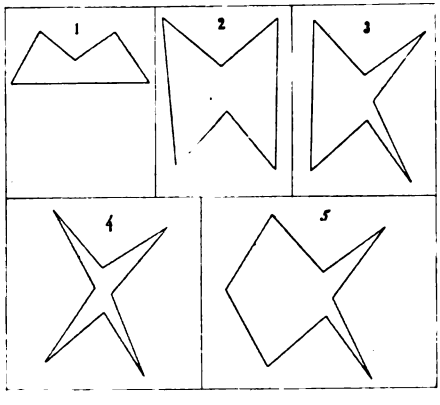

Figure 7

D. Final Conclusions on the Calculation of the Inductance of Single-turn Closed Aerials

Of the two methods of calculation proposed in Section B, the second is, naturally, to be preferred as the more simple. It gives in general the same degree of accuracy as the first. Thus, the general formula for the calculation of the inductance of single-turn closed aerials and, in general, plane figures having no reëntrant angles, the value of r/l being small, will be taken as

$$L = 2l\left(\log\frac{2l}{r} - a_k + \mu\delta\right) \qquad (31)$$

in which l is the perimeter of the figure r, radius of the wire; μ, magnetic permeability of the wire material; δ, a correction factor for frequency, determined in turn by formula (5) and the curve in Fig. 1; a_k, a constant, being a function of $l/\sqrt{s}$ and taken

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TABLE III

Point number	Type of figure	$\frac{l}{\sqrt{s}}$	a_k	Source
1	Circle	3.545	2.451	Formulas (4) and (32)
2	Regular octagon	3.641	2.561	Formulas (10a) and (36)
3	Regular hexagon	3.722	2.636	Formulas (9a) and (35)
4	Square	4.000	2.853	Formulas (6) and (33)
5	Equilateral triangle	4.559	3.197	Formulas (8a) and (34)
6	Equal-leg right triangle	4.828	3.331	Formulas (13) and (37)
7	Quadrilateral Fig. 3.	4.38	3.227	Formula (14)
8	Regular pentagon	3.812	2.712	Formula (37a)
9	Quadrilateral of Fig. 3a	4.395	3.091	Formula (37b)
10	Irregular pentagon Fig. 4	4.546	3.116	Exact solution for L taken with (31) to solve for a_k

from the curve in Fig. 6. The plotting of this curve becomes accordingly a matter of special interest.

The above pairs of values of a_k and $l/\sqrt{s}$ were used. The points are indicated in Fig. 6 by the same numbers as those here given.

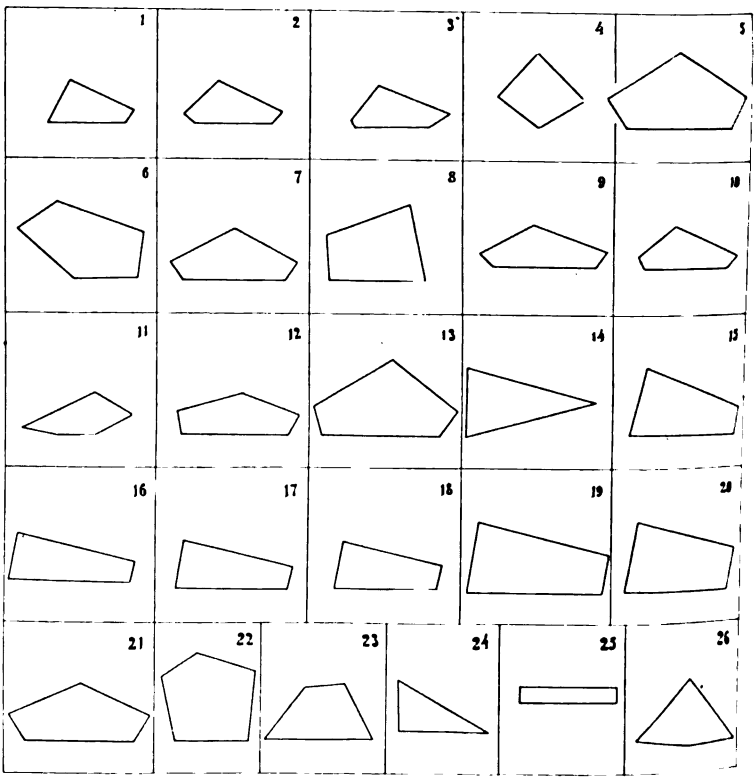

Figure 8

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Fig. 6 shows the curve plotted by the above method. The values of a_k for this curve up to $\frac{l}{\sqrt{s}}=4$ coincide completely with the theoretical values; further, up to $\frac{l}{\sqrt{s}}=4.828$ the curve is drawn through points giving sufficiently accurate values (better than 1 per cent). The part of this curve beyond the abscissa value $\frac{l}{\sqrt{s}}=4.8284$ is plotted by extrapolation, and is, of course, only a rough approximation. Table IV shows the results of a comparison and measured values[12] of the inductance for many antennas, the sizes and shapes of which are shown in Fig. 8.

TABLE IV

	Closed aerials		Perimeter	Area	$\frac{l}{\sqrt{s}}$	Radius of Wire	L measur.	L calcul.	ΔL per cent
1	The closed aerials of Lübertzy's radio receiving station near Moscow (Types 1920–1924)	Warsaw	285	3705	4.68	0.2	552000	546950	−0.92
2		Nauen	310	4373	4.68	0.2	606000	605800	−0.03
3		San-Paolo	315	4315	4.794	0.2	609000	604610	−0.72
4		Paris I	160	1593	4.00	0.2	307700	300160	−2.45
5		South	275	4525	4.08	0.2	565000	542620	−3.96
6		Tashkend I	304	4816	4.38	0.2	598000	594440	−0.6
7		S.-E.	335	5736	4.42	0.2	685000	670430	−1.56
8		Carnarvon I	220	2794	4.17	0.2	445000	421650	−5.25
9		Carnarvon II	375	5868	4.89	0.2	753000	729525	−3.12
10		Paris II	370	5038	5.2	0.2	740000	708300	−4.32
11		Tashkend II	350	5005	4.95	0.2	695000	673960	−3.02
12		America	385	6234	4.87	0.2	785000	751675	−4.24
13		West	330	4352	5.00	0.2	666000	630300	−5.35
14	The isosceles triangle		45.2	85	4.9	0.2	81000	75025	−7.37
15	The irregular quadrilateral		44.9	101	4.46	0.2	76500	76435	−0.085
16	The irregular quadrilateral		61.4	158	4.88	0.2	104000	105790	+1.72
17	The irregular quadrilateral		57.85	152	4.69	0.2	100000	100025	+0.02
18	The irregular quadrilateral		54.3	141	4.57	0.2	95000	93865	−1.19
19	The irregular quadrilateral		50.7	135	4.365	0.2	89000	88105	−1.00
20	The irregular quadrilateral		47.2	122	4.275	0.2	83000	81815	−1.43
21	The irregular pentagon		145	1175	4.23	0.2	275000	284750	−3.51
22	The irregular pentagon		52.12	187	3.81	0.2	98000	94775	−3.29
23	The irregular quadrilateral		45.94	115	4.28	0.2	75100	79385	+5.7
24	The right triangle		15.92	9.86	5.07	0.04	25500	25820	+1.25
25	The rectangle		70	150	5.71	0.1	121500	116970	−3.73
26	The Marconi d. f. aerial		97.5	330	5.37	0.1	164000	172535	+5.2

ΔL_m per cent = mean value for 26 examples = −1.39 per cent

The fact that on the average the calculated value of the inductance is smaller than the measured value in the above instances is probably to be attributed to the fact that the measured values include the inductance of the lead wires from the measuring apparatus to the aerial. This could readily account for the difference above.

[12] All measurements of inductance were made with a Siemens and Halske low-frequency bridge using standard coils of 10^5, 10^6, and 10^7 cm. by the same firm. The mean accuracy of the measurements was about 0.5 per cent for inside work and about one percent for outside work. Calculations were generally made with a 25 cm. slide-rule.

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Limits of Applicability of Formula (31)

The one approximation[13] used in deriving the expression (31) was the neglecting of terms of the order of r/l (first and higher powers), which are small in comparison with the main terms. It is just these quantities by which the formulas (1, 6, 8, 9, 10) and those forming a basis for deriving formula (31), i.e., formulas (32, 33, 34, 35, 36) differ. It follows that the limits of applicability of formula (31) are determined by the relative importance of values of the order of r/l and the remaining terms in the parenthesis of formula (31). Formulas (6, 8, 9, 10) for regular polygons differ from the approximate formulas (33, 34, 35, 36) in that to the expression for the inductance deduced from the fundamental formulas on which is based our curve (6), the quantity $2nr$ is to be added (in this, n is the number of sides of the polygon). In analogy to the square, formulas (6) and (33), it may be supposed that the accurate formulas should contain, in the factor in brackets, a term of the order of r/l, if we did not neglect quantities of the order of r/l to simplify the integration. This quantity must, similarly, figure in formula (31). The value of this correction for a figure of any shape remains as yet undetermined. For *regular figures* formula (31) may be probably made a closer approximation by introducing into the brackets the term $r \cdot \frac{n}{l}$. Then the

[13] Dr. Grover has been so kind as to check more in detail the assumption on which the derivation of formula (31) has been based, that the value of a_k is a function of $\frac{l}{\sqrt{s}}$ only. As a result of his analysis he has pointed out that in general the values of a_k given by the curve of Fig. 6 are only approximations and that the constant a_k although a function of $\frac{l}{\sqrt{s}}$ in every case is also dependent upon *the type of polygon* in question; that is, a_k is not the same for different types of figure, even though they have the same value of $\frac{l}{\sqrt{s}}$. Dr. Grover has plotted a_k as a function of $\frac{l}{\sqrt{s}}$ for triangles different types. The points all lie on the same curve, which is the same as that determined by points 5 and 6 in Figure 6. The curve calculated for rectangles, on the contrary, leaves this curve at the point 4, and runs below it, diverging gradually for greater values of $\frac{l}{\sqrt{s}}$.

In conclusion, however, Dr. Grover believes that the second method of calculating the inductance (formula 31) is capable of giving numerical values with an error not greater than a few per cent and because of its convenience is to be recommended as a valuable simplification for routine calculations of the inductance of irregular plane polygons of round wire.

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formula for all regular figures will read:

$$L=2l\left(\log\frac{2l}{r}-a_k+\mu\delta+r\frac{n}{l}\right) \tag{31a}$$

Taking the square as an example, the limits of applicability of formula (31) have been found, as well as the error resulting in using formula (31) instead of the exact formula (16) for the square. For various l/r ratios a correction in per cent is obtained (see

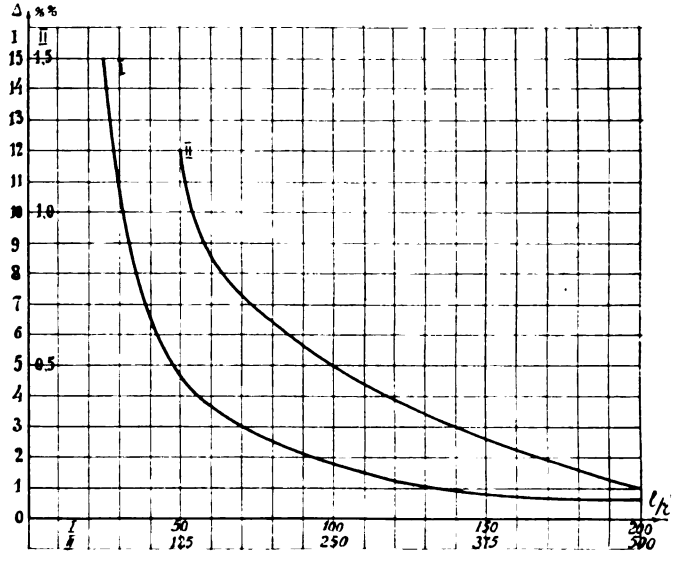

Figure 9

curves in Fig. 9). (The calculations have been made for the most disadvantageous case (radio frequency) when $\mu\delta=0$.) The value found from (31) must be augmented by just this quantity to obtain a more accurate expression for the inductance of a square of given dimensions. There is good reason to suppose that the value of this correction in per cent remains constant for all other figures. If so, our formula (31) with the addition of the correction Δ in per cent (from Fig. 9) will be true for all figures, all frequencies, and all values of l/r:

$$L=2l\left(\log\frac{2l}{r}-a_k+\mu\delta\right)(1+\Delta) \tag{31b}$$

This correction curve confirms the opinion we have expressed (Section B) that in all calculations of closed aerials ($l/r>500$)

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the correction in per cent may be neglected, the error not exceeding in the most disadvantageous conditions 0.1 per cent.

The advantage in using in calculations the simple formula (31) can be increased still more if we construct abacs for every term contained in the brackets of the expressions (31) or (31b).

(1) If the value of L is to be determined for high-frequencies only, the formula may be transformed into

$$L = 2l\left(\log\frac{2l}{r} - a_k\right) \tag{39}$$

By means of the abac (see Fig. 10) we find the value of $\log\frac{2l}{r} = f\left(\frac{l}{r}\right)$ and by means of curve (6) we determine $a_k = f\left(\frac{l}{\sqrt{s}}\right)$ Subtracting the second value from the first and multiplying the difference D by the quantity $2l$ (all in cm.) we obtain (also in cm.) the required value of the inductance.

(2) To find the inductance for low frequency ($\mu\delta$) we read in the nomogram (Fig. 11) the perimeter x, for given f (or λ) and r and obtain from the curve in Fig. 1 the corresponding value for $\delta = f(x)$. Multiplying the last by μ we add the product to the difference D found above. Multiplying the resulting sum by $2l$ we obtain the required value of the inductance at the given frequency.

(3) Lastly, when formula (31b) is applied to general calculations in electrical engineering or physics, for any possible value of the ratio l/r, a correction must be made; this is obtained from the curve in Fig. 9: $\Delta = f\left(\frac{l}{r}\right)$. The value of Δ (divided by 100) corresponding to the given l/r is to be added to 1 and the result found above in (1) or in (2) is to be multiplied by this sum.

(4) This formula is completely applicable to curvelinear figures (ellipse etc.); until now there have been no methods and formulas for these cases (excepting, of course, for the circle). It is likely that these formulas are applicable to the calculation of the inductance of power lines, etc., at least for one phase. The author is continuing work in this direction.

(5) As to the application of these formulas in the deduction of which the quasi-stationary condition is assumed to the case of high frequency currents in long wires (aerials) they are valuable to the cases of a normal working (sending or receiving) when the natural wavelength of the antenna has been increased to a length

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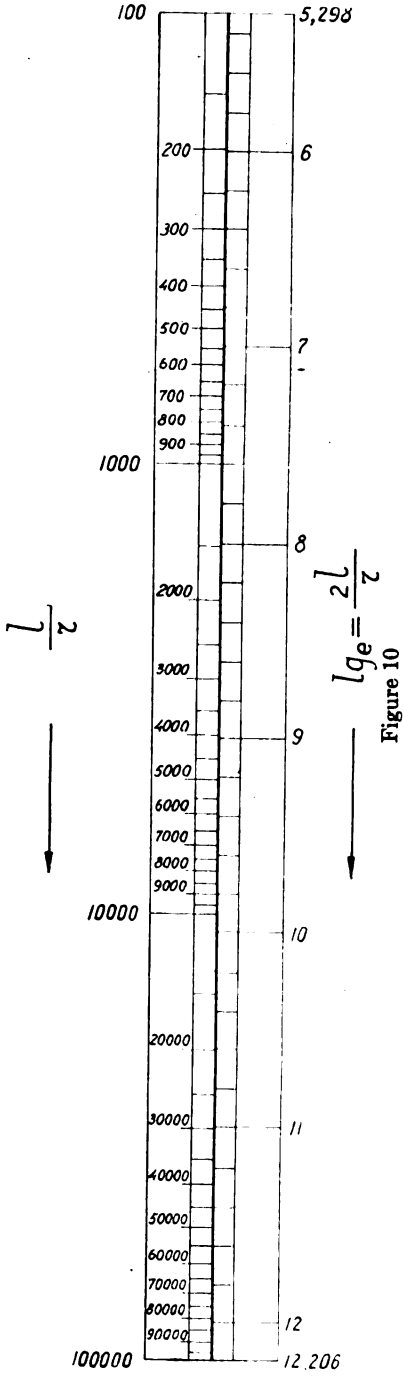

Figure 10

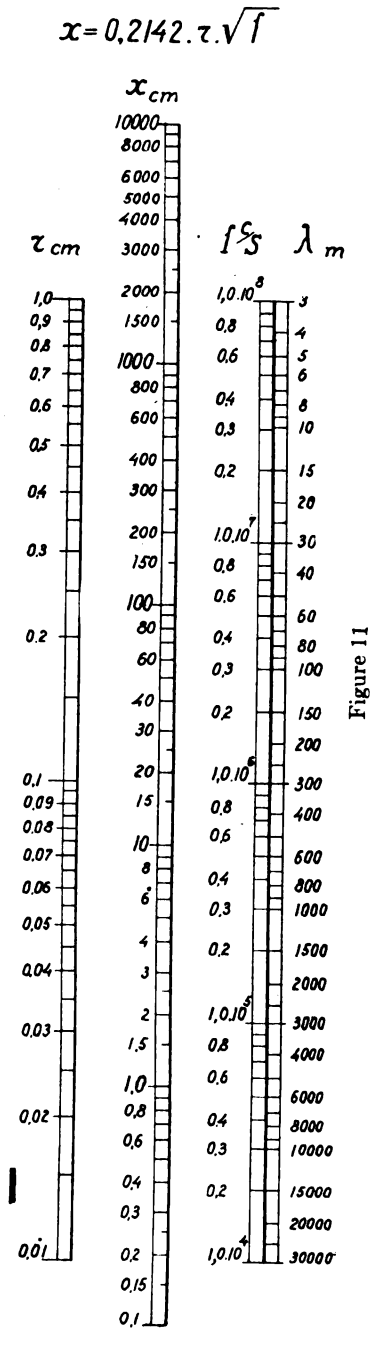

Figure 11

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more than twice as great by the insertion of coils or condensers. Otherwise our formulas give only the static inductance, from which one must find the effective (dynamic) inductance.

There was made one comparison between the effective (dynamic) value of L (measured with Behnken's method) and L calculated with the above formula (31). The isosceles triangle had the dimensions: $a=b=7.12$ m.; $c=9.88$ m.; the leads' length was 6 m. For wavelengths $l=250$ m. and $l=600$ m., the differences between L measur. and L calcul. were 2.82 per cent and 2.61 per cent respectively.

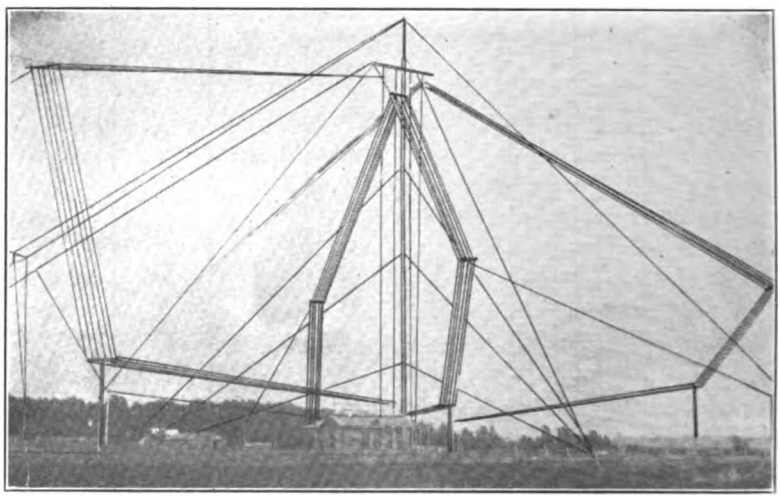

Figure 12—General View of Lubertzy's Mean Receiving Station (Construction of 1924)

In conclusion I wish to express my heartiest thanks to Professor M. V. Schuleikin, of Moscow, and to Professor F. W. Grover, of Washington, for helpful suggestions during this work. The latter has given assistance in the revision of the English copy of the manuscript.

The measurements were made in the Radio Department of the State Electrical Research Institute with the help of the Department's staff. To those who took part I offer my warmest thanks; especially I am obliged to M. E. Starik, electrical engineer, and N. K. Swistoff, student, who besides their work in the measurements afforded me very real help by making some of the calculations.

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SUMMARY

The article contains the first part of an investigation on the design and use of closed aerials (calculation of the inductance for single-turn aerials). Two methods are given for determining the inductance for closed aerials and, in general, plane figures of any shape, having no reëntrant angles. The first method is based on the author's supposition of the practical equivalence of the inductance of any figure and the inductance of a right triangle or rectangle, having area and perimeter equal to those of the given figure. The second method is a development of the basic assumption of the first, and results in the derivation of a simple formula for all possible figures, including curved ones.

Numerous tests of the correctness of both methods by practical measurements on closed aerials with a height of the point of support from 65 down to 2 meters have fully confirmed the author's conclusions.

Appendix

Dr. Grover has communicated the following formula for the inductance of any type of triangle, which we publish with his permission. The lengths of the sides of the triangle are taken to be a, b, and c.

$$L=2\Big[a\log\frac{2a}{r}+b\log\frac{2b}{r}+c\log\frac{2c}{r}-(a+b+c) \tag{40}$$

$$-(b+c)\operatorname{arc\,sin\,h}\frac{c^2+b^2-a^2}{v}-(a+b)\operatorname{arc\,sin\,h}\frac{a^2+b^2-c^2}{v}-$$

$$-(a+c)\operatorname{arc\,sin\,h}\frac{a^2+c^2-b^2}{v}\Big]$$

where $v^2=2(a^2c^2+a^2b^2+b^2c^2)-a^4-b^4-c^4$.

The same formula for the case of an isosceles triangle (the sides: a, a, and c) takes the form:

$$L=2\Big[2a\log\frac{2a}{r}+c\log\frac{2c}{r}-(2a+c)-2a\operatorname{arc\,sin\,h}\frac{2a^2-c^2}{c\sqrt{4a^2-c^2}}$$

$$-2(a+c)\operatorname{arc\,sin\,h}\frac{c}{\sqrt{4a^2-c^2}}\Big] \tag{41}$$

Finally for a right triangle this general formula (40) transforms itself in the following one:

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TABLE V

NN	Type of figure	Perimeter meters	Area of figure square meters	Radius of wire cm.	$\frac{l}{\sqrt{s}}$	L Measured	L calculated from formula (12)	L calculated from formula (43)	$\Delta L_{12}\%$	$\Delta L_{43}\%$
1	The right triangle ($a=3.9$ m., $b=17.7$ m., $c=18.1$ m.)	39.7	34.515	0.15	6.75	59100	52530	53195	−11.12	−10
2	The right triangle ($a=3.4$ m., $b=5.8$ m.)	15.92	9.86	0.04	5.07	25500	25970	25460	+ 1.84	− 0.157
3	The rectangle ($a=1$ m., $b=8$ m.)	18	8	0.1	6.36	25700	23365	23540	− 9.08	− 8.4
4	The rectangle ($a=5$ m., $b=30$ m.)	70	150	0.1	5.715	121500	115355	114850	− 5.05	− 5.4
5	The closed aerials of Lübertzy's radio-receiving station near Moscow: Karnarvon	375	5868	0.2	4.89	753000	739005	724650	− 1.86	− 3.75
6	Paris	370	5038	0.2	5.2	740000	709165	700200	− 4.17	− 5.4
7	Tashkent	350	5005	0.2	4.95	695000	681515	668500	− 1.94	− 3.8
8	America	385	6234	0.2	4.87	785000	762695	747750	− 2.84	− 4.75
9	West	330	4352	0.2	5.00	666000	635910	624200	− 4.52	− 6.25

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$$L=2\left[a\log\frac{2a}{r}+b\log\frac{2b}{r}+c\log\frac{2c}{r}-(a+b+c)\right.$$
$$\left.-(a+c)\operatorname{arc\,sin\,h}\frac{a}{b}-(b+c)\operatorname{arc\,sin\,h}\frac{b}{a}\right] \tag{42}$$

Passing from hyperbolic functions to usual geometrical, the last formula can be written in the form

$$L=a\left[\frac{a}{c}\log\frac{a(c-a)}{b(b+c)}+\log\frac{4a^2b^2}{r^2(c+a)^2}-2\right]$$
$$+b\left[\frac{b}{c}\log\frac{b(c-b)}{a(c+a)}+\log\frac{4a^2b^2}{r^2(c+b)^2}-2\right] \tag{43}$$
$$+c\left[\log\frac{4a^2b^2c^2}{r^2(a^2+ac)(b^2+bc)}-2\right]$$

In these formulas terms in $\frac{r}{a}$, $\frac{r}{b}$, and $\frac{r}{c}$ have been neglected. They do not include the internal linkages $\mu\delta$. Our earlier formula (12) gives values a little different from (43), the values being in general a little larger. Although a comparison of the inductances of the aerials shown in Table V makes clear that the inductances calculated by (12) are in good agreement with the measured values (at least for the larger aerials), the author believes that Dr. Grover's formulas (40) to (43) are a little more accurate, since the neglected quantities are smaller than those neglected in (12).

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PIEZO-ELECTRIC RESONANCE AND OSCILLATORY PHENOMENA WITH FLEXURAL VIBRATIONS IN QUARTZ PLATES

By

J. R. Harrison

(Wesleyan University, Middletown, Connecticut.)

The resonating and oscillatory effects of piezo-electric crystals were first described by Cady.[1] It is well-known that a rectangular plate cut from a quartz crystal having length l, breadth b, and thickness e, parallel respectively to the Y axis, Z (optic) axis and X (electric) axis has three fundamental natural frequencies as determined by these dimensions with a series of overtones for each. The length l of the crystal determines the frequency for the "transverse effect," the thickness e, that for the "longitudinal effect" and the breadth b, a third vibration frequency.[2] Each effect is due to a compressional wave traveling in one of the three respective dimensions of the crystal. In each case the electromagnetic wavelength corresponding to the frequency of the vibration is of the order of 110 meters per millimeter of crystal in the direction concerned.

In the course of some recent investigations with quartz plates it was found possible to excite plates into resonance at a frequency considerably lower than that of the transverse fundamental.[3] The transverse fundamental was the lowest frequency hitherto obtained in any quartz resonator as will be seen. This new piezo-electric resonance reaction has been found to be due to a flexural vibration in the length-breadth plane of the crystal, which is illustrated in Fig. 1, in which n, n represent the nodes which are $0.22\ l$ from the ends of the bar. When the electric field is applied in a manner herein explained, the stresses are such as to tend to make the crystal assume a shape as shown by the dotted lines. A reversal of the polarity of the applied field reverses the mechanical stresses so that the crystal edges will now be concave upward. An alternating electric field causes a rapid series of these concave upward and downward positions, the period of which is determined

[1] W. G. Cady, Proc. I. R. E., 10, pp. 83–114, 1922. J.O.S. and R.S.I., pp. 475–489, 1925.

[2] A. Hund, Proc. I. R. E., 14, pp. 455–456, 1926.

[3] Preliminary reports of this work have appeared in *The Physical Review*, 29, p. 366, 1927, and 29, p. 617. 1927.

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by the frequency of the applied electromotive-force. Resonance effects are obtained with this flexural mode of vibration exactly analogous to those obtained for the three simple modes due to compressional waves in the quartz. The response frequencies that have been obtained for several crystals are shown in Table I.

TABLE I

Crystal	Length	Breadth	Thickness	Flexure	Transverse Fundamental	Flexure	Transverse Fundamental
	mm.	mm.	mm.	kcs.	kcs.	meters	meters
E1	40.03	10.04	0.603	29.70	64.5	10100	4650
E2	40.03	10.05	0.617	30.0	67.8	10000	4430
E5	30.06	10.10	1.11	51.4	93.0	5850	3220
E6	30.06	10.10	1.12	50.9	92.2	5900	3250
W2	3.90	2.00	1.10	4,920.	6,990.	610	430
HA15	21.72	2.90	1.32	33.6	128.	8920	2350

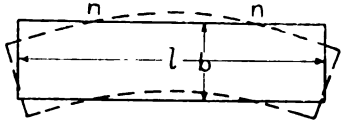

Figure 1—Flexural vibrations in the length-breadth plane showing nodes n, n

The frequency of a *thin* bar vibrating with any mode of flexural vibration is determined by the equation:[4]

$$N = \frac{m^2 k}{2\pi l^2}\sqrt{\frac{q}{d}} \tag{1}$$

where N is the frequency, k the radius of gyration, l the length, q Young's modulus, d density, and m a constant depending upon the mode of flexural vibration.

$$k = \sqrt{\frac{I}{M}} \tag{2}$$

where I is the moment of inertia of a cross section about a transverse axis through its center and M is the mass of the bar. Equation (1) holds accurately for flexural vibrations in the length-thickness plane so long as the thickness is small, but is equally applicable to flexural vibrations in the length-breadth plane for a bar whose breadth is small in comparison to its length. In this latter case the moment of inertia is taken about a transverse axis through its center:

$$I = \frac{1}{12}Mb^2 \tag{3}$$

[4] Barton, "Text-Book on Sound," p. 281 et. seq. or Horace Lamb, "Dynamical Theory of Sound," p. 122 et. seq.

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where b is the breadth of the bar. Substituting the values in (1) determined by equations (2) and (3), $m=4.73$ for the first mode of flexural vibration and the constants for quartz $d=2.654$ grams per cubic centimeter, $q=7.85\times10^{11}$,[5] the frequency for flexural vibrations in the length-breadth plane of a quartz bar is:

$$N=5.79\frac{b}{l^2}\times10^5 \text{ or} \tag{4}$$

$$\lambda=518\frac{l^2}{b} \tag{5}$$

where λ is the electromagnetic wavelength in meters corresponding to the frequency N, l and b are the length and breadth of the bar in centimeters, respectively.

Comparison of the observed frequency of flexural vibration with that computed from equation (4) shows a large discrepancy with bars in which the ratio of length to breadth is not large, just as would be expected. A series of tests were accordingly made to find the relation between the response frequency on the first flexural mode and the physical dimensions of the bar. Table II shows the results of a series of measurements taken on several plates of practically the same length and breadth but of different thicknesses. This conclusively shows that the response frequency of the crystal is independent of the thickness as it should be according to the theory.

In this Table II the transverse fundamental frequencies (longitudinal vibration) and the corresponding wavelengths have also been recorded for the eight plates. As would be expected from theory the thickness has not an appreciable effect.

TABLE II

Crystal	Length	Breadth	Thickness	Frequency		Wavelength	
				Flexural	Transverse Fundamental	Flexural	Transverse Fundamental
	mm.	mm.	mm.	kcs.	kcs.	meters	meters
A1	28.62	9.65	1.044	53.8	96.8	5580	3100
A2	28.58	9.65	1.058	53.3	96.5	5640	3120
A3	28.62	9.65	1.999	53.3	96.5	5630	3120
A4	28.62	9.66	2.015	52.6	95.6	5700	3140
A5	28.60	9.65	4.257	53.5	95.3	5620	3150
A6	28.64	9.66	4.256	52.6	95.0	5700	3160
A7	28.56	9.63	2.907	53.3	95.8	5640	3130
A8	28.64	9.63	3.557	52.6	96.8	5700	3100

[5] This value of Young's modulus was computed from the mean value of the number of meters per millimeter for the transverse fundamental of quartz taken from the results of the following observers: Mallet and Terry, *Wireless World*, **16**, pp. 631–636, 1925. Giebe and Scheibe, *Zeit. f. Physik*, **33**, pp. 335–344, 1925. Cady, Proc. I. R. E., **12**, pp. 805–816, 1924. Bureau of Standards, Letter Circular, No. 186.

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Measurements were made of the response frequency for the flexural vibration and the transverse fundamental as the length l of the crystal was diminished by grinding. Table III gives a typical set of tabulated results such as were obtained in this manner for several crystals. The observed results agree with those computed from equation (4) or (5) only when the ratio of length to breadth is large. The logarithmic plot of the length l of the crystal against the electromagnetic wavelength corresponding to the observed flexural response frequency of the crystal is shown in Fig. 2 by curve 1 for the results given in Table III. The slope of this curve 1 is less than would be expected from the theoretical

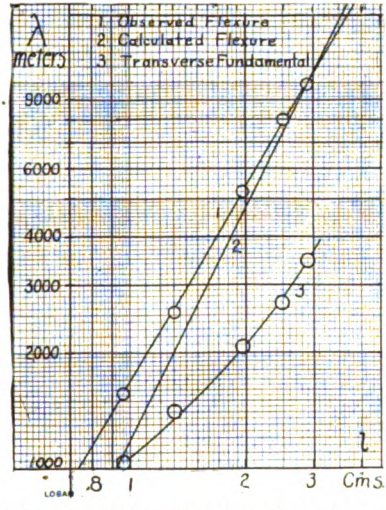

Figure 2—Logarithmic plot of length of crystal plate against the electromagnetic wavelength corresponding to the response frequency for the first mode of flexure and the transverse fundamental, the breadth remaining constant.

formula for flexural vibrations (5) the values computed from which are shown by curve No. 2 on the same graph. The slope of the curve 2 is 2 but curve No. 1 has a slope of only 1.7. This indicates that the exponent of the length l for these bars should be 1.7 instead of 2 as the theory indicates. Similar measurements were made on the variation of the flexural response frequency of a short bar or plate as the breadth b was diminished by grinding. Here again the numerical value of the exponent is less than that given by the flexural theory. From these results an approximate empirical formula has been developed which gives much better agreement with the observed flexural response frequencies of short bars than

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the theoretical formula (4). This formula for the electromagnetic wavelength corresponding to the response frequency for flexural vibrations in short bars is:

$$\lambda = k\frac{l^{1.7}}{b^{0.75}} \tag{6}$$

The value of k is nearly a constant for a series of measurements such as those given in Table 3 and is approximately equal to the constant in equation (5) for all but very short bars.

TABLE III

	Length l in cms.	Observed Flexure in meters	Calculated Flexure in meters	Observed Flexure in kilocycles	Calculated Flexure in kilocycles	Transverse Fundamental	
						meters	kcs.
1	2.89	9900	9850	30.4	30.4	3460	86.7
2	2.82	9540	9350	31.5	32.1	3200	93.7
3	2.67	8750	8390	34.3	35.8	2880	104.2
4	2.50	8000	7350	37.6	40.8	2680	112.0
5	2.37	7240	6610	41.5	45.5	2530	119
6	2.25	6630	5960	45.3	50.4	2400	125
7	1.97	5200	4560	57.7	65.8	2070	145
8	1.31	2535	2020	118.3	148.5	1410	213
9	0.96	1570	1080	191.0	278.0	1040	288

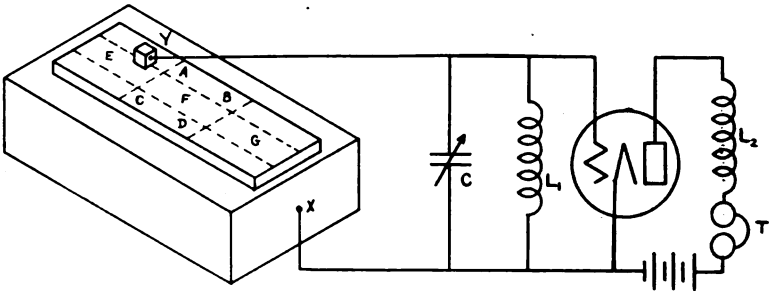

Figure 3—Simple oscillator circuit with large electrode X and small exploring electrode Y for locating regions of maximum resonance response in the piezo-electric crystal EG.

Resonance effects with the new low-frequency vibration were strongest when only half of the breadth of the quartz plate was enclosed by the metallic electrodes of the crystal mounting. When the whole crystal was enclosed by the mounting so that the applied electric field was practically uniform over the entire side of the plate the response at this frequency was found to be extremely faint and in many cases not audible in the phones. A simple vacuum-tube oscillator was used in these tests as shown in Fig. 3; the coils L_1 and L_2 were loosely coupled. Resonance conditions were found by noting the ringing sound or click of the crystal as the capacity C_1 was tuned through the critical frequency. In the following work a two-stage audio-frequency amplifier was

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used in connection with this oscillator circuit to make the method sensitive enough for the tests.

A study was made of the crystal response when very small exploring electrodes were placed at different points on the crystal. The crystal was placed on a metallic plate the dimensions of which were somewhat greater than those of the crystal. The exploring electrode Y (Fig. 3) was only $2\times2\times2$ millimeters square and was connected by a very fine wire to the oscillator circuit. The electrodes were connected to the filament and grid of the vacuum tube in the oscillator circuit as shown in Fig. 3. With this exploring

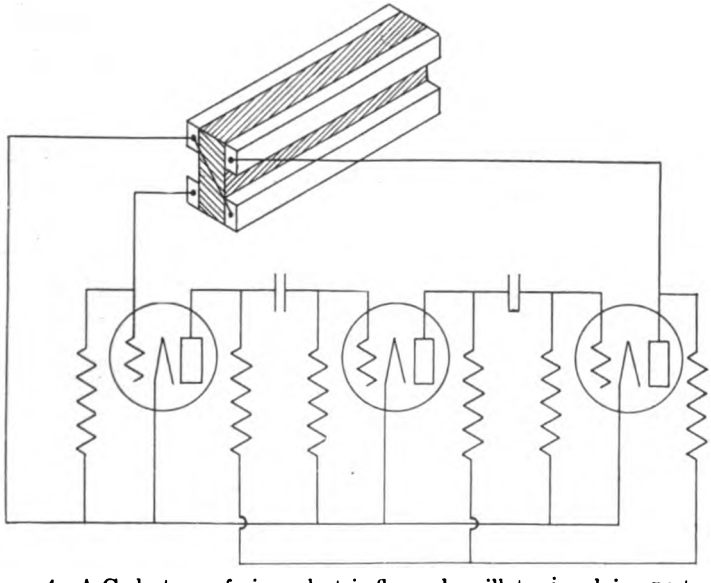

Figure 4—A Cady type of piezo-electric flexural oscillator involving no tuned electrical circuit.

electrode it was possible to explore the surface of the crystal and locate areas of maximum resonance response. These areas were indicated by the loudness of the click in the phones as the condenser in the vacuum-tube oscillator circuit was tuned through the flexural resonance frequency of the crystal. Maximum response regions were found near the upper and lower edges of the crystal, especially the central regions AB and CD as indicated in Fig. 3. When the exploring electrode was placed in the central part of the crystal at such points as E, F, or G the response was found to be very weak and in most cases inaudible. Two exploring electrodes were then connected in parallel and used on the upper side

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of the crystal, the lower electrode remained as in the previous test, with one exploring electrode placed in the region AB and the other at CD; no response was heard on tuning through the flexural resonance frequency. If both electrodes were placed together in either the region AB or CD the response was found just as would be expected.

A crystal mounting was then so designed that the electric field was applied to the two areas which were symmetrically disposed with respect to the longitudinal axis of the quartz plate and the polarities in the quartz were in opposite directions. Such a crystal mounting is shown in Figs. 4, 5, and 7. The four metallic electrodes A, B, C, and D shown in Fig. 7 are equal to the crystal in length and about one-third as wide. The upper and lower electrodes are separated by strips of insulating material. Electrode A is con-

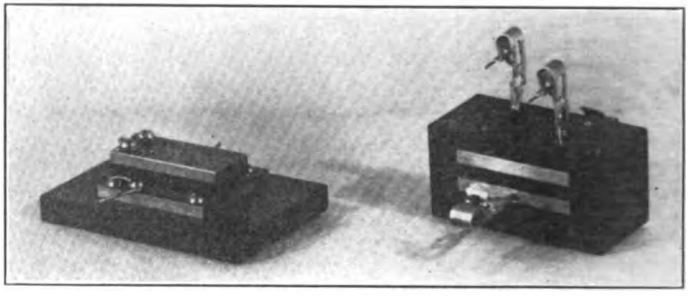

Figure 5—Two examples of the new type of crystal mounting with four electrodes for the first mode of flexural vibration.

nected to D, and B to C, and a binding post provided for each pair of electrodes. This same crystal mounting may be used to obtain the transverse fundamental and the longitudinal fundamental by connecting electrode A to B and C to D, so that they are effectively a single pair of plates. Fig. 5 shows crystal mountings such as were used in these tests. These were designed for plates of the same size and are identical except that one is completely enclosed to protect the crystal plate from dust and moisture.

The Piezo-Electric Oscillator

The crystal will also function as a piezo-electric oscillator on this flexural mode of vibration. Since the crystal mounting consists of four metallic plates, either a Cady or Pierce type of circuit may be used. One of the types of crystal oscillator circuit described

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by Cady[6] furnishes the necessary feedback of energy from output to input, to maintain sustained oscillations, through the vibrating piezo-electric crystal. To accomplish this the crystal mounting contains four electrodes or the equivalent, one pair of electrodes being connected to the input and the other pair to the output of a thermionic vacuum tube. The crystal mounting for the Pierce[7] type of oscillator requires but two electrodes or the equivalent connected to either grid and filament or grid and plate of the tube. The feedback in this case is produced entirely through the inter-electrode capacity of the thermionic tube unless a pickup or sensitizing coil is connected in series with the crystal. In this article crystal mountings are described having as many as twelve or more electrodes, but these are so interconnected that they act effectively as a mounting of two or four electrodes depending upon the type of circuit used. Fig. 4 illustrates this Cady type of circuit, as originally used, and involves neither inductance nor capacity, since the resistance-coupled amplifier insures sufficient energy to maintain the crystal oscillations. Two diagonally opposite electrodes A and D of the crystal mounting are connected to the filaments; the other electrodes B and C are connected to the output and input respectively.

With both Cady and Pierce type of circuit some difficulty is usually experienced in obtaining crystal oscillations if but a single vacuum tube is used. This trouble may be entirely overcome by introducing a pickup[8] or sensitizing coil in series with the crystal in the grid circuit. This method is illustrated in the Cady type of circuit by Fig. 6. The pickup coil L_1 should be loosely coupled to the inductance L_2 of the tuned plate circuit. The coupling between L_1 and L_2 is made as loose as possible to avoid cracking the crystal with excessive power, and also to avoid setting up oscillations not of crystal frequency, but due to the electrical constants of the circuit alone. Experiments indicate that under these conditions the frequency of the oscillator seems to be as constant as for the other modes of vibration. With the Pierce type of circuit, Fig. 7, one pair of diagonally opposite electrodes of the crystal mounting are connected to the grid of the vacuum tube and the other pair to the filament. As before in the Cady type of oscillator circuit a pickup coil L_1 is used in series with the crystal in the grid circuit. This type of circuit seems to be most promising for future develop-

[6] W. G. Cady, Proc. I. R. E., **10**, p. 112 et seq. 1922.
[7] G. W. Pierce, Proc., Amer. Acad. of Arts & Sci., **59**, pp. 79–106, 1923.
[8] A. Hinderlich, *Experimental Wireless*, **4**, pp. 29–41, 1927.

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ment since the output power developed is considerably greater than that obtainable from the Cady type of circuit and arcing between the plates of the crystal mounting is not experienced at normal plate voltages as is found to be the case with the former circuit. A plate of quartz 30×10×1 mm. vibrating flexurally at 60 kilocycles was used in the Pierce circuit as a power oscillator. A type UX-210 tube was used with 400 volts on the plate. The observed output power was about 0.5 watts, which considering the low frequency compares favorably with the output from crystals vibrating at high frequencies.[9]

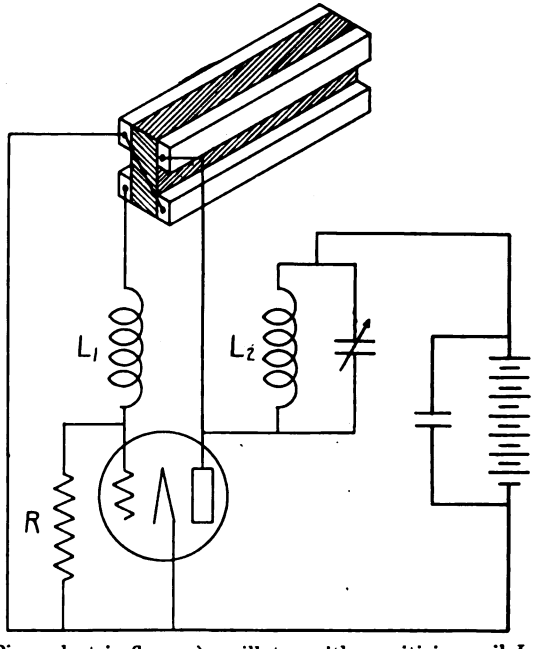

Figure 6—Piezo-electric flexural oscillator with sensitizing coil L_1 in the grid circuit.

Higher Modes of Flexure

Other modes of flexural vibration of higher frequency and in the length-breadth plane of the crystal have also been obtained. Resonance effects have been found and regions of maximum resonance response determined just as described above with small exploring electrodes. To obtain oscillations on these higher modes of flexural vibration, it is desirable to use a crystal mounting that

[9] A. Crossley, Proc. I. R. E., 15, pp. 9–36, 1927.

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applies the electric fields in such directions that the resultant stresses in the crystal are a maximum. For the second mode of flexural vibration the mechanical stresses are such as to make the crystal assume the shape shown in Fig. 8. The regions n are nodes.

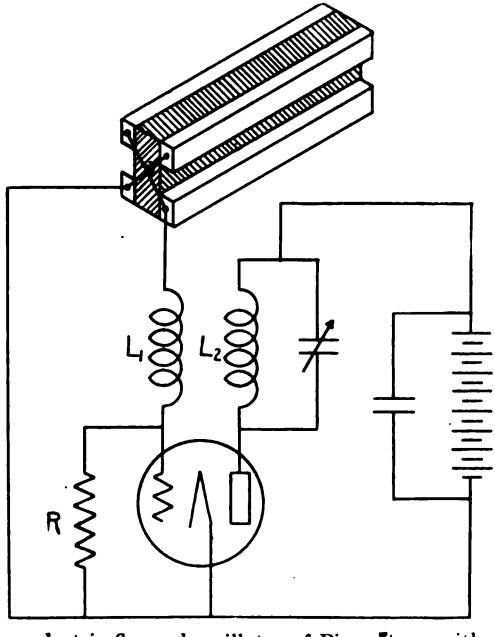

Figure 7—Piezo-electric flexural oscillator of Pierce type with sensitizing coil L_1 in grid circuit.

two of which are 0.12 of the length l from the ends and the third is centrally located along l according to theory and roughly confirmed by actual observation. The existence of these nodes is quite strikingly illustrated by sprinkling a little lycopodium powder on the long edge of the vibrating crystal. The lycopodium

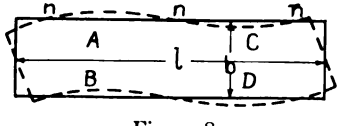

Figure 8

will pile up in the nodal regions n showing them to be regions of minimum motion. When the regions A and D in the crystal are in a state of tension regions B and C will be in compression and vice versa. Accordingly the electric field applied to the regions A and D should be opposite in direction to that applied at B and C.

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Crystal mountings which apply the electric fields in this manner are illustrated in Figs. 9 and 10. The crystal mounting (Fig. 9) consists of eight metallic electrodes *S*, *T*, *U*, *V*, *W*, *X*, *Y*, and *Z*

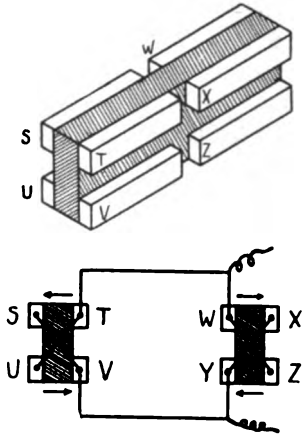

Figure 9—The upper diagram shows the new crystal mounting of eight electrodes for the second mode of flexural vibration. The lower diagram indicates the method of connecting electrodes. The arrows indicate the relative direction of the electric field in various portions of the crystal.

which are slightly less than one-half the crystal in length and about one-third as wide. These electrodes are arranged along and flush with each long edge of the crystal as shown. The electrodes *S* and *V* are connected together by a wire as are likewise the follow-

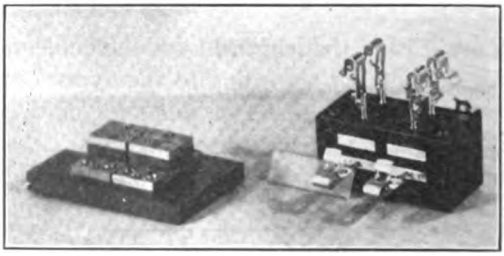

Figure 10—Two examples of crystal mountings with eight electrodes for the second mode of flexural vibration. By properly interconnecting the electrodes this may also be used to excite the first mode of flexural vibration and various other frequencies by means of the transverse and longitudinal effects.

ing electrodes: *T* and *U*, *X* and *Y*, and *W* and *Z*. The pair *S* and *V* are now connected to the pair *X* and *Y* and likewise the pair *T* and *U* are connected to the pair *W* and *Z*. These connections are illustrated in the lower part of Fig. 9 where *both* ends of the crystal and

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the mounting are shown. The direction of the electric field for each pair of electrodes is indicated by arrows. In like manner a crystal mounting may be designed for the third mode of flexural vibration, but four additional electrodes will be required, making twelve in all. This type of crystal mounting is indispensable for crystal oscillators, but is not necessary for simple resonance effects as has been pointed out before. Table 4 gives a typical set of data for the frequencies of the first and second modes of flexure and of the transverse fundamental as the length is diminished by grinding. The equation for the second mode of flexure is identical with that for the first mode (equations (5) and (6)) except that a new constant m must be introduced (see equation (1)). These observations, Table 4, indicate that the frequency increases for all modes as the length is diminished. The frequency for the transverse fundamental increases linearly with diminishing breadth. The frequency for the flexural modes increases at a much greater rate. The second mode of flexure becomes of higher frequency than the transverse fundamental in short bars.

The only reason for using these flexural vibrations is to extend the range of response frequencies obtainable from a crystal plate on the low frequency side. The limit of utility is then reached when a given mode of flexural vibration is of higher frequency than the transverse fundamental. The transverse fundamental was the lowest frequency hitherto obtained from a crystal. The response frequency for the third mode of flexural vibration is usually higher than that of the transverse fundamental except in the case of long narrow bars, since then all the flexural response frequencies are correspondingly lower.

TABLE IV

Length	Frequency First Mode of Flexure	Frequency Second Mode of Flexure	Frequency Transverse Fundamental	Wavelength First Mode of Flexure	Wavelength Second Mode of Flexure	Wavelength Transverse Fundamental
mm.	kcs.	kcs.	kcs.	meters	meters	meters
28.9	30.4	72.0	86.7	9900	4160	3460
28.2	31.5	75.0	93.7	9540	4000	3200
26.7	34.3	81.1	104.	8750	3700	2880
25.0	37.6	92.0	112.	8000	3260	2675
22.5	45.3	109.	125.	6630	2750	2400
19.7	57.7	135.	145.	5200	2220	2070

Luminous Phenomena

Giebe and Scheibe[10] have recently shown that the resonance state of a piezo-electric crystal may be made directly visible under

[10] Giebe and Scheibe, *Zeit. f. Physik*, **33**, pp. 335–344, 1925. *Elektrotech. Zeit.*, **47**, pp. 380–85, 1926. H. Kroncke, *Wireless World*, **17**, p. 896, 1925.

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certain conditions by the luminous glow emitted. With electrodes that cover the full length and breadth of the plate, this luminous glow is observed in the air gap between the crystal and mounting, when resonance conditions are reached in a partially evacuated chamber. Dye[11] has more recently pointed out that the voltage gradient in the quartz is enormously increased when an air gap exists between the crystals and the electrodes. The voltage across the air gap may be 30 times the voltage applied to the electrodes. Under these conditions and with the crystal placed in a suitable vacuum a glow discharge may be expected.

Before making tests for luminous effects with flexural vibrations, observations were made according to Giebe and Scheibe of the crystal glow for the transverse fundamental and its overtones. In all these tests the plate was placed flush with one side of the crystal mounting, so that but one air gap existed between the two. The luminous glow for the transverse fundamental effect is of maximum intensity at the center of the plate along the length l; the intensity diminishes somewhat on both sides of this region. The total region of luminosity extends over about one-third the length l of the crystal, varying somewhat with air gap, applied electromotive-force and probably other conditions. With a suitable crystal mounting consisting of four electrodes (the Cady type) as described by Giebe and Scheibe two regions of maximum luminous intensity were found for the first overtone of the transverse fundamental. The other overtones may also be indicated in a similar manner.

The luminous glow emitted by crystals can also be observed at the flexural vibration frequency and under certain conditions presents striking peculiarities which still await explanation. With a single pair of electrodes covering the whole of both sides of the the crystal no luminous glow appears at the flexural vibration frequency as would be expected, for the crystal does not vibrate flexurally under these conditions. Using the new crystal mounting for flexural vibrations as described in this paper, Fig. 5, a luminous glow was observed in the air gap between the crystal and electrodes which very much resembled that found for the transverse fundamental. With a simple crystal mounting of two electrodes which cover only one-third of the breadth and approximately the full length (see Fig. 11), a peculiar striated luminous glow appears on the upper edge XY of the crystal and not, as might be expected, in the air gap. At higher pressures when the glow first appears,

[11] D. W. Dye, *Proc.*, Phys. Soc. of London, **38**, p. 453, 1926.

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it is always as described above in the form of beads or striations running perpendicular to the length l, that is, parallel to the direction of the applied electric field. As many as seven beads have been observed with some crystals, the form of which is approximately represented on the upper edge of the crystal in Fig. 11. Many photographs of this effect have been made, but the results are not sufficiently good to warrant direct reproduction.

To make the beads visible, the frequency of the applied electromotive force must be carefully tuned to crystal resonance. When the applied frequency is slightly detuned from crystal resonance the glow diminishes in intensity and the beads are

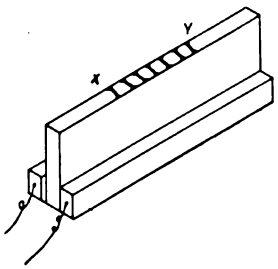

Figure 11—Piezo-electric crystal with simple electrodes covering only one-third of the breadth. The appearance of the beads or striated luminous glow is indicated on the upper edge XY of the crystal.

replaced by a faint but uniform luminosity. The beads will only be visible for a short period of time unless the oscillator is free from small transient shifts in frequency. As the vacuum is raised in the crystal chamber the beads diminish in number and become more diffuse and indistinct.

The striated appearance of the glow discharge may be due to some peculiar and as yet unexplained effects in the mechanical vibration at the center of the plate. It is to be noted that the amplitude of vibration of the crystal is considerably less with the mounting (Fig. 11) than would be the case with the special mounting for flexural vibrations. The striated glow is only present when a single pair of electrodes is used and appears on the edge most remote from the electrodes. With the four electrode mounting the peculiar mechanical effect may still be present, but only the usual type of discharge between electrodes and crystal is then observed as previously described.

When the luminous effect is to be used as a visual indicator of frequency, best results will be obtained with a residual gas of

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helium or neon in the crystal chamber. Similar results are obtained with air, but the luminous glow is not so intense.

The luminous effect is sometimes obtained even at atmospheric pressure when the applied voltage is sufficiently great. There is considerable danger of damaging the crystal under these conditions because of excessive mechanical strain.

Finally, the author wishes to thank Professor W. G. Cady, under whose direction the problem was undertaken, for his kind advice and constant cooperation.

Summary

A method of obtaining flexural vibrations from piezo-electric quartz plates together with the equations for flexural vibrations in quartz is given. Crystal mountings for obtaining best results with the first, second, and higher modes of flexural vibration are developed. The use of the crystal at flexural frequencies in the oscillator circuits of Cady and Pierce, and the luminous phenomena obtained when resonating in a partially evacuated chamber with different types of mountings, are also described. Without using unusually long plates the range of frequencies obtainable is extended into the audible range.

Among the problems still to be solved are the temperature coefficient of frequency and logarithmic decrement for the various modes of flexural vibration as well as constancy of frequency under variations of load and constants of the electric circuit. Investigations are still in progress, of which a report will be communicated later.

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Discussions on

TELEPHONE COMMUNICATION OVER HIGH POWER LINES BY HIGH FREQUENCY CURRENTS* (BODDIE)

R. D. Duncan, Jr.:† This paper is one of the very few concerning high-frequency communication over power lines which have been published in this country. Mr. Boddie's statement and analysis of the fundamental problems involved and of the requirements of equipment for this service are clear and comprehensive. It is with the section of the paper on "Attenuation Constant at High Frequencies" and those following that this discussion is concerned.

In the sections referred to comparison is made between measured and computed values of the inductance, capacity, and resistance per unit length of overhead high voltage power lines, for different physical arrangement of conductors and for frequencies in the neighborhood of 60 kilocycles. A good agreement between measured and computed values of the linear inductance and capacity for full-metallic systems was obtained; there was considerable difference, however, between similar values of the high-frequency resistance of the lines, the measured values in every case being much greater than those computed from well-known skin effect formulas. For full metallic circuits the differences were within the limits 181 to 294 per cent; for circuits with ground return much greater differences were obtained.

It is suggested in the paper that for the full metallic circuits, these differences were due in part to the presence of nearby parallel conductors in which currents were induced. In the line-ground system they were attributed to ground-return resistance. Neither of these effects is taken into account by the theory upon which the computed values of resistance were based.

The manner of behavior of overhead lines at high frequencies, when free in space or when surrounded by other lines, and the advantages or disadvantages of full metallic versus line-ground return carrier circuits, have undoubtedly received the consideration at one time or other of every engineer engaged in this particular branch of radio. In the field of wired radio broadcasting

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* PROC., Vol. 15, 7, Page 559, July 1927.
† Wired Radio, Inc., New York City.

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wherein high-frequency currents are superimposed upon the electric light and power lines of a community, every advantage has been with full-metallic operation. It requires less transmitting power than the line-ground system and is not subject to fading of signals, line noises and interference from high-power space radio as is the latter system. Full-metallic carrier operation produces the minimum inductive interference in nearby overhead lines (for example, telephone lines) and space radio receivers, antenna or loop types which may be in operation near the lines. A great advantage of full-metallic operation is that it lends itself readily to the use of polyphase carrier which is referred to later in this discussion. To investigate the carrier properties of overhead lines the necessary measurements were carried through for both full-metallic and line-ground systems to enable the derivation of the unit resistance, inductance, capacity, and leakage conductance, over the range of frequencies normally employed in carrier operation. It is believed that a statement of these results and a comparison thereof with those obtained by Mr. Boddie will not be without interest and will assist in explaining some of the differences between theory and experiment noted in the paper under discussion.

Mr. Boddie derived values of linear resistance by measuring the characteristic impedance Z_0 of the lines, and the transmitted and received currents with the lines terminated in an impedance, Z_0. The resistance R[1] is computed from the approximate expression

$$R = 2 \cdot \alpha \cdot Z_0 \tag{1}$$

which is the first term of the more general expression

$$\alpha = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}} \tag{2}$$

under the assumption

$$Z_0 = \sqrt{\frac{L}{C}} \tag{3}$$

In the expressions α, L, C, and G are respectively the attenuation constant, inductance, capacity, and leakage conductance.[1] The attenuation constant α is related to the ratio of transmitted to

[1] Here taken as per loop mile for full-metallic and per wire mile for ground-return circuits.

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received current of the lines of length l, when terminated in Z_0 through the usual expression

$$\frac{I_T}{I_R} = \epsilon^{\alpha l} \tag{4}$$

In utilizing relations (1), (3), and (4) for deriving the resistance, the leakage conductance is implicitly assumed to be zero. This assumption may or may not be valid for highly insulated high-voltage lines. It was not found to be true for low voltage power lines. Certain types of insulators while entirely satisfactory at commercial power frequencies, exhibit losses at carrier frequencies and are unsatisfactory. The possibility of such losses occurring even with high voltage insulators, thereby requiring the consideration of the second member of expression (2) was suggested by one of the members during the discussion of Mr. Boddie's paper. The results of measurements quoted herewith show that at least for some types of power lines the leakage conductance is appreciable and cannot be neglected. They also show, however, that including the second term in expression (2) does not account for all of the difference between the measured and computed values of line resistance, and that factors other than ground resistance cause a large portion of the line losses in a line-ground system.

In explanation of these statements the following experimental data[2] is submitted which gives the characteristic impedance, and the inductance, capacity, resistance, leakage conductance, attenuation, and wavelength constants, per loop mile for full-metallic circuits and per wire mile for line-ground circuits, for aerial conductors of average height above ground of 30 feet and spacing (horizontal) between conductors of 42 inches. The lines are No. 6 copper, diameter 0.162 inch, approximately one-half mile in length. Each conductor has standard weather-proof insulation and was supported by porcelain pin insulators such as are normally employed in 2300 volt construction. The lines were strung on the same pole with other lines and for a portion of their length jointly occupied the same cross arm with other power lines. In general their construction was typical and the reaction due to the proximity of adjacent lines may be considered typical of what is encountered in overhead city construction.

[2] The line measurements and computations were made by Messrs. S. Isler, S. A. Barone, G. C. Salmons, and C. E. Bohner, of the engineering staff of Wired Radio, Inc.

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TABLE I

FULL-METALLIC SYSTEM

1 Frequency	2 Char. Impedance Z_0	3 Inductance L	4 Capacity C	5 Leakage Conductance G	6 Resistance R	7 Attenuation Constant α	8 Wavelength Constant β	9 Resistance (Computed)	10 Inductance (Computed)	11 Capacity (Computed)
K.c. per sec.	Ohms (1)	Henrys (2)	μf (2)	$\mu mhos$ (2)	Ohms (2)	Hyp. Radians (2)	Cir. Radians (2)	Ohms (2)	Henrys (2)	μf (2)
30	698	0.00423	0.00867	14.28	13.1	0.0143	1.133	11.4	0.00413	
40	703	0.00422	0.00853	15.77	19.1	0.0189	1.506	13.2	0.00412	0.00715
50	706	0.00426	0.00853	12.29	24.0	0.0214	1.895	14.7	0.00411	
60	706	0.00424	0.00849	10.00	30.0	0.0259	2.260	16.1	0.00410	
70	697	0.00420	0.00863	7.74	36.7	0.0290	2.647	17.4	0.00410	

(1) Phase angle of the order of 20 minutes negative.
(2) Per loop mile.

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The experimental procedure followed, consisted in measuring the input resistance and reactance[3] for the two conditions of the far ends of the lines, open-circuited and short-circuited upon themselves. Thence by well established methods[4] the unit constants were derived. No special difficulty has been encountered in applying this method at carrier frequencies.

The data in Table I is for the full-metallic system, and in Table II, for a line-ground system. Columns 1 to 8 inclusive in both tables give measured values, and columns 9, 10, and 11 give values computed from the dimensions of the system.

In Table I for the full-metallic system the characteristic impedance Z_0 is seen to be practically independent of the frequency with an average value of 700 ohms, and virtually non-reactive. The measured and computed values of inductance L differed by 3 per cent, and of capacity C by approximately 16 per cent. The ratio of measured to computed resistance varied from 1.15 at 30 kc. to 2.11 at 70 kc. The difference in inductance values is too small to have any significance. The differences in the capacity and resistance values, however, are large and in the case of the latter are of the order of those observed by Mr. Boddie.

For the line-ground system of Table II the characteristic impedance decreased with increasing frequency and, as compared with the full metallic system, was smaller and showed a higher negative phase angle. The inductance and capacity per wire mile on the average showed an increase of 24 per cent and 11 per cent respectively over the computed values. The computed skin effect resistance may be taken as one-half of that of the full metallic system, given in column 9, Table I; the ratios of measured to computed resistance for 30 and 60 kilocycles are respectively 10.17 and 13.25 or much higher than with the full-metallic system. It is noted that the leakage conductance of the full metallic system is less, that is, the effective leakage resistance is higher, than with the line-ground system.

The increase in unit resistance and capacity of the lines in full-metallic operation is undoubtedly due to the proximity of lines running parallel and close to those measured. By virtue of their coupling with the lines in question an added resistance is introduced and the total interwire capacity increased. In the line-ground system because of the more widely distributed char-

[3] By a method similar to that outlined by Mr. Boddie. See also "Recent Attainments in Wired Radio," *Journal of Franklin Institute*, Jan. 1921.

[4] Pernot, "Electrical Phenomena in Parallel Conductors," Vol. 1, Chapt. VIII.

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TABLE II

LINE-GROUND SYSTEM

1 Frequency	2 Char. Impedance Z_0	3 Inductance L	4 Capacity C	5 Leakage Conductance G	6 Resistance R	7 Attenuation Constant α	8 Wavelength Constant β	9 Resistance (Computed)	10 Inductance (Computed)	11 Capacity (Computed)
K.c. per sec.	Ohms (1)	Henrys (2)	μf (2)	$\mu mhos.$ (2)	Ohms (2)	Hyp. Radians (2)	Cir. Radians (2)	Ohms (2)	Henrys (2) (3)	μf (2)
30	608	0.00387	0.0105	32.5	58.0	0.0576	1.203	5.7		
40	566	0.00359	0.0112	35.0	67.0	0.0692	1.594	6.6	0.002606	0.00983
50	529	0.00303	0.0108	54.0	86.0	0.0956	1.801	7.3		
60	523	0.00326	0.0119	18.8	106.0	0.1064	2.354	8.0		

(1) Phase angle of the order of 2 deg. negative.
(2) Per wire mile.
(3) Neglecting skin effect.

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acter of the electric field surrounding the conductor it is to be expected that the proximity effect of adjacent conductors would be very pronounced, and hence would contribute largely to the change in the line constants observed. Mr. Boddie has ascribed the entire change to ground-return resistance, and in accordance with the theory developed by Carson[5] the resistive component of the ground-return impedance of an overhead line is by no means negligible. Considerable experimental work at high frequencies with overhead lines under congested pole line conditions, however, has indicated that nearby lines may greatly increase the effective attenuation of a distributed circuit, and it is believed that in line-ground systems the change in line constants due to the proximity of other conductors is of equal importance to the ground-return effect.

Two other sources of loss which should be mentioned are those due to electromagnetic radiation and, in the case of the paper under discussion, the steel core of the stranded conductor. In accordance with what little is known concerning radiation from long distributed systems the resistance so introduced at these carrier frequencies appears to be negligible. The close agreement between the measured and computed values of inductance quoted by Mr. Boddie would indicate that at carrier frequencies the current penetration into the conductor does not reach the steel core.

Another point of interest in comparing Tables I and II is the lower characteristic impedance of the line-ground system which, however, has the much higher attenuation. That is, for the same applied voltage a higher input current is obtained, but less of it is effective in reaching the receiving end of the lines. Both of these features have been observed experimentally.

Because of the growing importance of wired radio on all types of line wire systems it is suggested than an experimental research into the high-frequency transmission characteristics of aerial conductors, both free in space and in the presence of other conductors in specified arrangements, is a problem well worth consideration. An experimental check upon existing theory would be provided and information of great value made available.

In concluding this discussion it is desired to direct attention to the advantages of employing polyphase carrier on polyphase power systems. In particular when communicating over a three-

[5] Carson, *B. S. T. J.* Vol. V, No. 4, page 539, 1926.

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phase power line, three-phase high-frequency energy, properly modulated would be impressed on the lines and all phases thereof utilized at both the transmitting and receiving terminals. A power transmission system is thus established similar in every respect to the normal power system and obeying the same laws but operating at a much higher order frequency. In wired radio broadcasting on polyphase systems where power for lighting is supplied from all phases extending from a substation, it is necessary to distribute high frequency current *uniformly* over all phases in order to obtain a uniform signal level. Single phase carrier has not provided this uniformity whereas polyphase carrier has successfully met all requirements.[6] In the general field of radio it is interesting to note that three-phase high frequency provides a means for effecting carrier-frequency suppression[7] and duplex radio telephone operation.[8] It is believed that polyphase carrier will play an important part in future methods of high-frequency communication.

Alexander Nyman:† *A considerable portion of the above paper is devoted to the calculation of antenna efficiency for transmission of signals to power lines. As a result of these calculations the author comes to the conclusion that the advantage of condenser coupling as compared with antenna coupling is negligible since the antenna efficiency from 97½ per cent to 94 per cent can always be secured.

On further investigation of this conclusion it appears that the transmitting set to which this antenna or condenser is connected was not taken into consideration in determining the efficiency, and yet it is this transmitting set in connection with the coupling system which really determines the efficiency of the whole operating scheme.

To illustrate it more clearly, consider the following:

The paper shows that the current through the power line is about one-quarter of the current through the antenna. As the antenna current must be supplied by the transmitting set, probably by some tuned means such as an inductance, it is quite evident that the losses in that inductance as well as in the antenna must be taken into account. Assuming the values of capacity of the antenna, which are given by the author as 0.003 μf., and an operat-

[6] U. S. Patent No. 1578881.
[7] U. S. Patent No. 1560505.
[8] U. S. Patent No. 1591025.
*Original Manuscript received by the Institute, September 21, 1927.
† Consulting Engineer, Dubilier Condenser Corporation.

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ing frequency of 50,000 cycles, an inductance of about 9.4 millihenrys will be necessary, with a reactance of 1060 ohms at that frequency. Since with the best possible design it is difficult to get a power factor less than 1 per cent on an inductance coil of this magnitude and this frequency, the resistance value would be about 10 ohms and with 1 ampere current the additional losses would be about 10 watts, which must be added to the transmitter power of 27 watts.

Thus the efficiency of the whole system would be now

$$\frac{27}{27+10} = 73 \text{ per cent}$$

If on the other hand instead of antenna a condenser of the same capacity (0.003 μfds) were used, the current supplied by the inductance coil would be exactly equal to the current put into the line, that is, 0.26 amperes, and the losses in that inductance coil instead of 10 watts, would be only 0.68 watts, giving a total efficiency of

$$\frac{27}{27+0.68} = 97.5 \text{ per cent}$$

These values of efficiency, of course, do not take into account the actual antenna or condenser efficiency; they are simply the values of efficiency as determined by the inductance coil. If as the author claims he can secure an efficiency of 94 per cent to 97½ per cent of the antenna, the above figure of 73 per cent should be multiplied by this antenna efficiency. On the other hand, with condenser coupling the power factor of the condenser is a fraction of 1 per cent and hence the efficiency of the condenser coupling device is more than 99 per cent.

In determining the effective resistance of the antenna in order to establish its efficiency, I further note that the author has apparently disregarded entirely the ground resistance, stating however that these ground losses are likely to be quite appreciable. It is incorrect to state that these "ground losses are not chargeable directly to the mechanism of coupling, but to the existing conditions" since under other conditions such as condenser coupling these ground losses disappear, or under the worst circumstances are reduced 16 times on account of practically 4 times smaller current. To make this clear, assume a resistance in the ground of 5 ohms. In case of antenna coupling, this would produce a loss

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of 5 watts and a reduction of efficiency as claimed by the writer from 97½ per cent to

$$\frac{27}{27.76+5}=82.5 \text{ per cent}$$

giving a total efficiency of

$$82.5 \text{ per cent} \times 73 \text{ per cent} = 60 \text{ per cent}$$

On the other hand with a condenser coupling and assuming as the worst case a ground return with the same resistance of 5 ohms, the current is only 0.26 amperes and the corresponding loss 0.34 watts. Thus, the efficiency instead of 97.5 per cent will be changed to

$$\frac{27}{27+0.68+0.34}=96.5 \text{ per cent}$$

These figures would give quite a different conclusion from that of the author, showing that the condenser coupling apart from the considerable simplification, in connection, and the improvement on the receiving circuit, would also give an efficiency of 96.5 per cent under worse circumstances, as against 60 per cent on the antenna coupling, or, in other words, the loss with condenser coupling is about 10 times smaller than in antenna coupling.

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Review of Current Literature

Prepared by

STUART BALLANTINE

Boonton - - - - - New Jersey

FURTHER COMMUNICATION ON THE PROPAGATION OF SHORT-WAVES.*

[BY

E. QUÄCK

(Berlin)

(*Zeitschrift für Hochfrequenztechnik*, **30**, 41, Aug. 1927)

In an earlier communication the occurrence of doubled-signals in the propagation of short waves was reported.[1] For the purpose of calculating the height of the Heaviside layer it was assumed in that paper that the waves were propagated with the velocity of light. According to more recent work,[2] however, it appears that

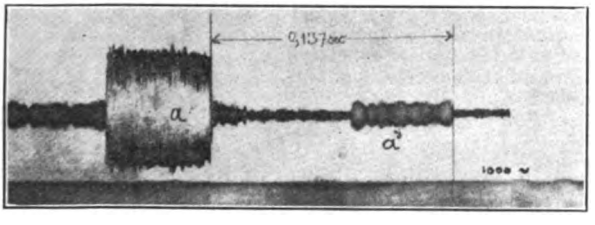

Figure 1—Doubled signals from transmitter SPU, Rio (15.5 m.) *a* direct signal; *a'* same signal which has made an additional circuit of the earth. Photographed May 31, 1927; 2135 E. M. T.

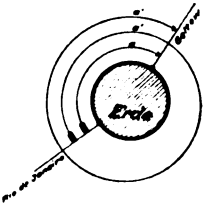

Figure 2—Schematic diagram of the paths taken by the signals shown in Fig. 1.

the velocity of propagation is smaller than the light-velocity; therefore the computed height of the Heaviside-layer will be correspondingly diminished. Consequently, a new determination of the height of the layer should be attempted on the supposition that the wave velocity is dependent upon the index of refraction and the frequency. Repeated observations at Geltow of the signals

* Original Manuscript received by the Institute, October 3, 1927.

[1] E. Quäck, *Zeit. für Hochfrequenztechnik*, **28**, 177, 1926: PROC. Inst. Radio Engs. **15**, 341; 1927.

[2] A. H. Taylor and E. O. Hulburt, *Phys. Rev.*, **27**, 198, 1926. H. Lassen, *Zeit. für Hochfrequenztechnik*, **28**, 109, 139, 1926.

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radiated at Nauen have yielded values between 0.137 and 0.139 sec. for the time difference between the direct signal and the signal which encircles the earth.

It has been impossible to discover any difference between the transmitters *AGC* ($\lambda = 18.22$ *m.*) and *AGA* ($\lambda = 14.9$ *m.*) at Nauen with respect to the time-lag of their direct and doubled signals. It is also remarkable that the doubling in the case of the Nauen transmitter *AGC* could be observed at night (22:00 European Mean Time.) These observations have been supplemented by those of several other observers, notably Dr. Moegel and Mr. Wiesner.

In May the Brazilian Radiotelegraph Company, Rio de Janiero, placed in operation the short-wave beam transmitter

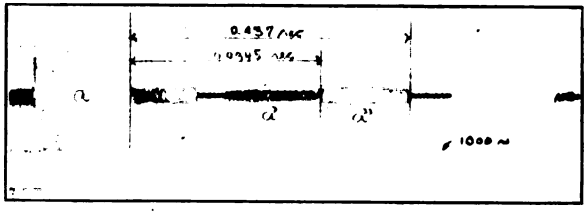

Figure 3—Tripled signal from WIK, Rocky Point (21.45 m.) *a* direct signal; *a'* secondary signal which has gone the other way around the earth; *a''* tertiary signal taking direction of direct signal but making an additional circuit of the earth. Photographed April 10, 1927; 0220 E. M. T.

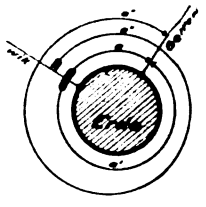

Figure 4—Schematic diagram of paths of signals *a*, *a'*, *a''* shown in Figure 3.

SPU, wavelength 15.66 m., and antenna power of 10 k. w. Strong doubled-signals from this transmitter have been observed from 1300 to 1800 G.M.T. and also from 2045 and 2330 G.M.T. During the latter period time differences between direct and doubled-signals were observed which correspond to a path of the secondary signal around the earth in the same direction as the direct signal, but making a complete additional circulation. The time-difference of this complete circuit amounts to about 0.1365 to 0.140 second. An oscillographically-recorded doubled-signal of this type is shown in Fig. 1; in Fig. 2 are shown the different paths taken by the signals. The direct signal *a* has the path *a* from Rio to Geltow; the doubled-signal *a'* takes the path *a'* from Rio over Geltow finally arriving at Geltow after completing the circuit of the earth. The doubled-signal thus traverses a path compared with that of the direct signal which corresponds to the observed time-difference of 0.137 second.

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Fig. 3 illustrates a tripled signal, the possibility of which is brought about by the wave taking a direct path, a path the other way around the earth, and a path of the type just described.

Observations were made on June 10th of this year at about 02:20 E.M.T. on the non-directional transmission of the North American short-wave transmitter *WIK* ($\lambda = 21.45$ m.). From many oscillographic records, of which Fig. 3 is a specimen, the following observations have been made: The direct signal *a* is followed after 0.0945 sec. by the secondary signal *a'* which has taken the other path around the earth to Geltow. After this appears the tertiary signal *a''* which starts out with the direct signal, passes over Geltow, and makes a complete circuit of the earth in this

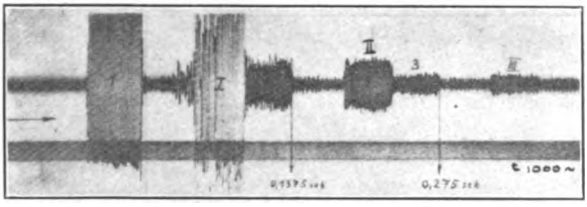

Figure 5—Double circulation around the earth. Tripled signal from SPU (15.5 m.), Rio. *1* and *I* direct signals, *2* and *II* signals in direction of direct signal but with additional circuit about earth, *3* and *III* in direction of direct signal but with two additional circulations. Photographed June 27, 1927, E. M. T.

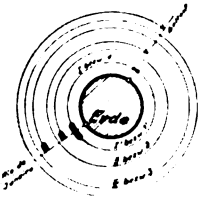

Figure 6—Schematic diagram of paths taken by signals shown in Figure 5.

direction previous to its arrival at Geltow after a time interval of 0.137 to 0.138 second. The amplitude of *a''* is greater than that of *a'* in spite of the longer path. These paths are represented schematically in Fig. 4.

Evidence that the wave may encircle the earth several times is furnished by Fig. 5. This record was obtained with the signals from the Brazilian station *SPU* previously mentioned. This station emits every second two dots spaced about 1/20 second apart which are photographed as the direct signals 1 and *I*. After 0.1375 sec. the secondary signals 2 and *II* are recorded after their complete circuit around the earth in the direction of the direct signals; after another 0.1375 sec. (or total of 0.275 sec.) the tertiary signals 3 and *III* are registered; these signals have thus twice encircled the earth. The paths are again schematically shown in Fig. 6.

Fig. 7 contains evidence of further signal multiplication, although perhaps not quite so clear and certain. In addition to the

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previously described doubled-signals the signals marked *I'* and 1' are discernible corresponding to waves which have taken the reverse path around the earth as shown in Fig. 6. The remarkable thing about this is that inasmuch as Rio is equipped with directional (beam) antenna system one would not expect to find recordable signals radiated in this direction.[3]

It appears as if this multiple circulation about the earth occurs with especial facility when the great circle containing the sending and receiving points lies essentially in the zone of twilight. This is especially the case in the Rio observations where this type of signal was nightly encountered from 21:00 to 22:00 E.M.T., at which time the great circle lay directly in the twilight. In daytime

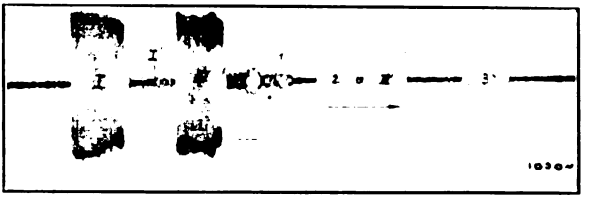

Figure 7—Quadrupled-signal from SPU, Rio (15.5 m.)

there occurred only the ordinary doubled-signals, corresponding to the reverse path around the earth. Up to the present time doubled-signals have been observed in the region of wavelengths extending from 14 to 34 meters. It is astonishing how much energy is still retained by the waves after several circulations around the earth, and it is certainly to be expected that further circulations occur in addition to those here recorded. For the practical use of short-waves, however, there must be found ways and means for avoiding the disturbances of operation occasioned by these "ghost" signals. Their systematic observation will aid in this, as well as contribute to the explanation of the propagation of waves of this type.

S. B.

[3] [This observation seems to be in agreement with the ideas of L. Bouthillon (*Comptes Rendus*, **184**, 190, 1924), who contends that inasmuch as the rays most effective in establishing long range communication are those making a small angle with the zenith and that beam projectors of the type of the Marconi Company at Bodmin radiate less directionally at such angles, the efficacy of such systems becomes less and less from the viewpoint of wave concentration as the distance increases. On the other hand, according to Hulburt's theory waves of this length are totally reflected only if their zenith angles are greater than 60–70 degrees, which would appear to conflict with Bouthillon's small zenith angles. Inasmuch as the precise mode of propagation of Quäck's ghost waves is still very much in doubt it does not yet seem possible to decide between these rival views.—S. B.]

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RÉSUMÉ OF PROF. G. W. O. HOWE'S CRITIQUE OF HERR QUÄCK'S COMPUTATIONS

In connection with the extraordinarily interesting and important observations of Herr Quäck on the propagation of short waves the comments of our English contemporary Prof. G. W. O. Howe, contained in an article entitled "Phase and Group-Velocities in an Ionized Medium,"[1] are of interest. A brief résumé of Howe's criticism follows:

Herr Quäck computed the apparent height of the great circle path of the ray corresponding to the oscillographically recorded "double-signal" by multiplying the time difference, as measured upon the oscillogram, by the velocity of light. This turned out to be 182 kms., although with the most recent value of the time interval given by Quäck (0.138 sec., *vide* the above article) the value of 205 kms. appears more accurate. In any event this is considerably greater than the ionic refraction theory and other experimental evidence as to the ionization at various heights would seem to indicate. Prof. Howe suggests that the divergence may be explained by the difference between the group and phase-velocities.

In works on optical theory it is shown that in the case of propagation in a dispersive medium, i.e., one in which the wave (phase) velocity is a function of the frequency, the velocity of a homogeneous group of such waves, measured let us suppose from the head of the group, will be given by:[2]

$$v_g = \frac{v}{1 - \frac{n}{v}\frac{dv}{dn}} \tag{1}$$

Now in the case of an ionized medium in which dispersion is caused by motion of the electrons, the wave-velocity is given by:

$$v = \frac{c}{\sqrt{1 - \frac{Ne^2c^2}{\pi m n^2}}} \tag{2}$$

the influence of collisions and of the earth's magnetic field not being considered. Hence the group-velocity is:

$$v_g = c\sqrt{1 - (Ne^2c^2)/(\pi m n^2)} \tag{3}$$

[1] *Exp. Wireless*, **4**, 259, May 1927.

[2] See T. H. Havelock; "Propagation of Disturbances in Dispersive Media," *Cambridge Tract*, No. 17, p. 3.

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For the group-velocity to differ by 1 per cent from the velocity of light, N, the number of electrons per cc, would have to be 5.6×10^4; for a difference of 5 per cent $N=28\times10^4$. These are quite reasonable values for N. Appleton and Barnett deduced for certain conditions a minimum value of N of 100,000 and Appleton has recently deduced a minimum value of 2.5×10^6. (More recent computations based upon radio transmission measurements yield values of N ranging from $3\text{–}6\times10^5$).

In view of this, Quäck's value of 182 km. for the height of the Heaviside layer would be reduced to 116.5 kms., for a velocity 1 per cent less than that of light, and to 51 kms. for a velocity 2 per cent less.

Prof. Howe suggests that the result of the experiment is therefore not to determine the height at which the ray travels, but rather to determine from the assumed height the value of electron density in the medium through which the ray travels. N varies but little for widely different assumptions as to the height; an accurate determination of the time taken to encircle the globe would therefore decide within narrow limits the electron density of the medium. Since there are no insuperable difficulties connected with the accurate measurement of the time-differences the method would appear to be a most sensitive and important one for the investigation of this fundamental quantity.

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BOOK REVIEW

Principles of Radio Communication by JOHN H. MORECROFT,[1] assisted by A. PINTO[2] and W. A. CURRY.[3] SECOND EDITION, 1001 Pages, Published by JOHN WILEY & SONS, New York. Price $7.50.

Morecroft's first edition came out just before the broadcasting cyclone struck. For a time many radio men referred to that first edition as their radio bible. But the broadcasting cyclone changed the relative importance of radio matters and sped up radio to such an extent that radio men have been asking why Morecroft did not get out an up-to-date edition. They have been asking that for the past three or four years.

The second edition is out now. The advertising sheets announcing that edition have been sent out, presumably to all members of the Institute; at least, one has been sent to me. A glance at this advertising might give one the impression that the book is chiefly about the old spark type of radio, because about the most conspicuous thing in that advertising is two cuts of one of the neat little spark transmitters which, according to my recollection, Emil J. Simon designed nearly fifteen years ago. They were good spark sets, but they are apt to give that advertising a through ticket to the waste basket, for who wants to buy a book about spark sets now. That is unfortunate because the book is very much more up-to-date than those conspicuous cuts imply.

Chapter V is headed "Spark Telegraphy." That chapter includes 77 of the 1001 pages in the book, and 27 of those 77 pages are about receiving circuits, detectors, head phones, amount of power in sound waves, audibility, and the characteristics of crystal rectifiers. Therefore it will be seen that spark telegraphy is only a minor part of this second edition. The first edition contained twelve more pages on that subject, although the total number of pages in the first edition was less than the total number in this second edition.

Chapter VI is the largest chapter in the book. That chapter takes up 240 pages or nearly one-fourth of the book, and its title

Original Manuscript received by the Institute, October 24, 1927.

[1] Professor of Electrical Engineering, Columbia University; past president of the Institute of Radio Engineers.

[2] Electrical Engineer, Otis Elevator Co.

[3] Assistant Professor of Electrical Engineering, Columbia University.

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is "Vacuum Tubes and their Operation." Sparks characterized the past and vacuum tubes characterize the present in radio. From those chapters it will be seen that the authors have tried to make this book as comprehensive and modern as a book can be made.

Presumably to give more room for recent developments, the authors have not only eliminated part of the chapter on spark telegraphy and put more pages in this edition than in the first edition, but they have eliminated chapters X and XII on "Wavemeters and their Use" and "Radio Experiments," which were in the first edition. I believe they should have put in and improved the chapter on "Radio Experiments," because it not only showed how experiments are carried on but included problems to be solved. Such a chapter is valuable to people who are learning about radio and a new group comes along every year to learn about radio.

This edition is divided into ten chapters. Each chapter has a title and is further divided by sub-titles. The titles and sub-titles appear in the table of contents and at the tops of the pages. Also, the sub-titles are included in the index at the back which occupies over twelve pages. This arrangement helps the user to find what he wants. The titles of the chapters are: Fundamentals, Ideas and Laws; Resistance—Inductance—Capacity—Shielding; Laws of Oscillating Circuits; General View of Radio Communication; Spark Telegraphy: Vacuum Tubes and Their Operation in Typical Circuits; Continuous-wave Telegraphy; Radio Telephony; Antennae and Radiation; and Amplifiers.

Broadcast transmitters, receivers and loud speakers are uppermost in many minds at present. These subjects are treated in the chapter on Radio Telephony. Receivers are also treated, for example, in the chapters on Spark Telegraphy and Amplifiers.

The volume is well illustrated on almost every page, some pages carrying several illustrations. Sketches, photographs, circuit diagrams, curves and, best of all, numerous oscillograms are used to show what happens in one, two, or three circuits at one time.

Not only do the authors consider many phases of radio subjects, but they have included numerous references at the bottom of pages, to special papers on those subjects.

The text begins with a discussion of the "Nature of Electricity" taking up "Electrons" as the next subject. The text ends with the diagram and description of an amplifier equipped with transformer, rectifier and filters to take the filament and 350-volt plate power from a 60-cycle line. That is an indication of how modern and comprehensive this book is. It is a big book, too big to describe well enough in this review.

R. H. Marriott

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued October 18, to November 15, 1927

By

JOHN B. BRADY

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,645,231—ELECTROMAGNETIC SOUND REPRODUCER—FREDERICK DIETRICH, of New York, N. Y., and WILLIAM H. GERNS, East Orange, N. J. Filed March 5, 1926, issued October 11, 1927. Assigned to Brandes Laboratories, Inc.

1,645,483—ELECTROMAGNETIC DRIVER—C. A. BRIGHAM and W. H. GERNS of East Orange, N. J. Filed August 22, 1925, issued October 18, 1927. Assigned to Brandes Laboratories, Inc.

1,646,292—METHOD OF DIRECTING LIGHT RAYS UPON LIGHT SENSITIVE ELEMENTS—CLINTON W. HOUGH, Boonville, N. Y. Filed Oct. 22, 1925, issued Oct. 18, 1927. Assigned to Wired Radio, Inc.

1,645,523—CRYSTAL DETECTOR OF ELECTRIC CURRENTS—H. M. DOWSETT, Colchester, England. Filed Dec. 18, 1920 issued Oct. 18, 1927. Assigned to Radio Corp. of America.

1,645,542—CIRCUIT ARRANGEMENT FOR HIGH FREQUENCY SENDING STATIONS—M. OSNOS, Berlin, Germany. Filed Dec. 27, 1922, issued Oct. 18, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie.

1,645,560—WIRELESS TELEGRAPH RECEIVING APPARATUS—R. A. WEAGANT, New York, N. Y. Filed April 14, 1920, issued Oct. 18, 1927. Assigned to Radio Corp. of America.

1,645,850—REGULATOR—F. S. BERNHARD, Brooklyn, N. Y. Filed Nov. 22, 1923, issued Oct. 18, 1927. Assigned to Western Electric Co.

1,645,904—ELECTRON DISCHARGE DEVICE—P. GAVIN, National Sanatorium, Tenn. Filed Aug. 5, 1925, issed Oct. 18, 1927.

1,646,152—RADIO CIRCUIT TESTING INSTRUMENT—J. O. KLEBER, Pittsburgh, Pa. Filed May 20, 1926, issued Oct. 18, 1927.

1,646,236—VARIABLE CONDENSER—W. DUBILIER, New York, N. Y. Filed Feb. 14, 1924, issued Oct. 18, 1927. Assigned to Dubilier Condenser Corp.

1,646,443—COMPENSATOR FOR RADIO COMPASSES—W. B. BURGESS, Boston, Mass. Filed Mar. 11, 1921, issued Oct. 25, 1927.

1,646,471—STATION RECORDING MEANS FOR RADIO RECEIVING DEVICES—A. ZILLGER, Narberth, Pa. Filed Dec. 9, 1924, issued Oct. 25, 1927. Assigned to Music Master Corp.

1,646,517—KEY CIRCUIT FOR TUBE SENDERS—W. ZELETSKI, Berlin, Germany. Filed Mar. 26, 1923, issued Oct. 25, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie M. B. S.

1,646,626—VACUUM TUBE—H. E. METCALF, San Leandro, Calif. Filed July 23, 1924, issued Oct. 25, 1927. Assigned to The Magnavox Company, of Oakland, Calif.

1,646,633—PROCESS FOR EXHAUSTING DETECTOR TUBES—R. T. ST. JAMES Chicago, Ill. Filed April 10, 1922, issued Oct. 25, 1927. Assigned to Chicago Miniature, Lamp Works.

1,646,663—CONDENSER—R. C. ROSE, Osceola, Arkansas. Filed Oct. 27, 1926, issued Oct. 25, 1927.

1,646,707—ACOUSTIC HORNS—W. R. RESPESS, St. George, N. Y. Filed (original) Dec. 8, 1923; divisional filed Sept. 30, 1924, issued Oct. 25, 1927. Assigned to Brandes Laboratories, Inc.

1,647,238—ELECTRON DISCHARGE DEVICE—E. H. MANTHORNE, Bayside, N. Y Filed June 18, 1925, issued Nov. 1, 1927. Assigned to Bell Telephone Laboratories, Inc.

1,647,283—ARRANGEMENT FOR IMPROVING SHORT WAVE RADIATION INTO SPACE—A. ESAU, Jena, Germany. Filed Oct. 8, 1926, issued Nov. 1, 1927.

1,647,290—ANTENNA SWITCHING RELAY—A. HADDOCK, East Orange, N. J. Filed June 22, 1922, issued Nov. 1, 1927. Assigned to Western Electric Co., Inc.

1,647,349—RADIO SIGNALING APPARATUS—H. T. FRIIS, Red Bank, N. J. Filed, June 9, 1925, issued Nov. 1, 1927. Assigned to Western Electric Co. Inc.

1,646,292—MOUNTING FOR ACOUSTIC HORNS—DAVID H. MOSS, Brooklyn, N. Y. Filed Feb. 13, 1924, issued Nov. 1, 1927. Assigned to Brandes Laboratories, Inc.

1,647,364—ELECTRONIC DEVICE—B. F. JANCKE, Brooklyn, N. Y. Filed Oct. 8, 1925, issued Nov. 1, 1927. Assigned to Manhattan Electrical Supply Co.

1,647,474—VARIABLE PATHWAY—F. W. SEYNOUR, Flushing, N. Y. Filed Oct. 25, 1925, issued Nov. 1, 1927.

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1,647,609—HETERODYNE RECEIVING SYSTEM—W. F. COTTER, Buffalo, N. Y. Filed Feb. 16, 1925, issued Nov. 1, 1927. Assigned to Federal Telephone Mfg. Corp.

1,647,617—ELECTRON DISCHARGE DEVICE—T. R. GRIFFITH, Dover, N. J. Filed July 7, 1926 issued Nov. 1, 1927. Assigned to Bell Telephone Laboratories, Inc.

1,647,994—ELECTRON DISCHARGE DEVICE—W. G. HOUSKEEPER, of New York, N. Y. Filed October 27, 1922, issued November 8, 1927. Assigned to Western Electric Co., Inc.

1,647,996—ELECTRON DISCHARGE DEVICE—F. L. HUNTER, Jr., of Towaco, and SYLVESTER W. CROWLEY, of Weehawken, N. J. Filed January 17, 1924, issued November 8, 1927. Assigned to DeDorest Radio Telephone & Telegraph Co.

1,648,183—METHOD AND APPARATUS FOR CONDUCTING CURRENT—K. H. KINGDON, and IRVING LANGMUIR, of Schenectady, New York. Filed December 21, 1922, issued November 8, 1927. Assigned to General Electric Co.

1,648,428—MACHINE FOR TESTING ELECTROMAGNETIC SOUND REPRODUCERS—RAY G. STACY, East Orange, N. J. Filed March 18, 1927, issued Nov. 8, 1927. Assigned to Brandes Laboratories, Inc.

1,648,293—ELECTRIC DISCHARGE DEVICE—ERNEST E. CHARLTON, of Schenectady, N. Y. Filed original Nov. 29, 1922. Renewed September 23, 1927, issued November 8, 1927. Assigned to General Electric Co.

1,648,312—ELECTRON DISCHARGE DEVICE—K. H. KINGDON and IRVING LANGMUIR, of Schenectady, N. Y. Filed November 6, 1923, issued November 8, 1927. Assigned to General Electric Co.

1,648,458—ELECTRON DISCHARGE DEVICE AND METHOD OF OPERATING THE SAME—G. M. J. MACKAY and ERNEST E. CHARLTON, of Schenectady, N. Y. Filed August 27, 1926, issued November 8, 1927. Assigned to General Electric Co.

1,648,521—RADIO RECEIVING SET—A. WIKSTROM, of Boston, Mass. Filed November 10, 1924, issued November 8, 1927. Assigned to Flash Radio Corp.

1,648,592—WIRELESS APPARATUS—L. R. RUOFF, of Syracuse, N. Y. Filed January 11, 1923, issued November 8, 1927. Assigned one-fourth to Howard P. Denison and one-fourth to Eugene A. Thompson, of Syracuse, N. Y.

1,648,682—VARYING FREQUENCY SYSTEM OF RADIOSIGNALING—J. H. HAMMOND, Jr., of Gloucester, Mass. Filed June 28, 1923, issued November 8, 1927.

1,648,835—TWO-WAY RADIO COMMUNICATION SYSTEM—P. WARE, of New York, N. Y. Filed September 4, 1920, issued November 8, 1927. Assigned to Ware Radio, Inc.

1,648,689—CONDENSER TRANSMITTER—AUGUST HUND, Bethesda, Maryland. Filed Apr. 10, 1926, issued Nov. 1, 1927. Assigned to Wired Radio, Inc.

1,648,941—MANUFACTURE OF OXIDE-COATED CATHODES—WILLIAM BENJAMIN GERO and GEORGE WILSON HALLOCK, Bloomfield, N. J. Filed July 16, 1926, issued Nov. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co..

1,648,958—EVACUATED DEVICE AND METHOD OF EXHAUST—RALPH E. MYERS, East Orange, N. J. Filed May 5, 1923, issued Nov. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,649,016—OLIVER E. BUCKLEY, Maplewood, N. J. Filed Dec. 13, 1917 (original) renewed Apr. 9, 1927, issued Nov. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,649,122—OSCILLATION GENERATOR—SIEGMUND LOEWE, Berlin, Germany. Filed Sept. 2, 1921, issued Nov. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,649,131—TUNING ARRANGEMENT FOR RADIO COMMUNICATION—C. SCHWARZ, Charlottenburg, near Berlin, Germany. Filed Oct. 31, 1921, issued Nov. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,649,391—PROTECTIVE SYSTEM FOR HOT CATHODE DEVICES—R. L. DAVIS, Wilkinsburg, Pa. Filed June 23, 1925, issued Nov. 15, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,649,489—VACUUM TUBE—H. J ROUND, London, England. Filed July 9, 1921 issued Nov. 15, 1927. Assigned to Radio Corp. of America.

1,649,499—RECEIVING ARRANGEMENT FOR WIRELESS TELEGRAPHY—G. G. VON ARCO, Berlin, Germany. Filed April 30, 1923, issued Nov. 15, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie.

1,649,510—WIRELESS INSTALLATION ON VEHICLES SUCH AS AUTOMOBILES—N. H. CLOUGH, London, England. Filed Oct. 23, 1923, issued Nov. 15, 1927. Assigned to Radio Corp. of America.

1,649,727—AERIAL FOR RADIO RECEPTION—C. F. PEITZMAN, Grimes, Iowa. Filed May 5, 1926, issued Oct. 15, 1927.

1,649,778—MEANS FOR AND METHOD OF CHANGING THE INTENSITY OF SIGNALS IN RADIO DYNAMIC RECEIVING SYSTEMS—J. H. HAMMOND, Jr. and E. L. CHAFFEE, of Gloucester, Mass. and Belmont, Mass. respectively. Filed (original) June 27, 1917, renewed May 31, 1924, issued Nov. 15, 1927. Chaffee assigned to Hammond.

1,649,257—SUSPENSION FOR ARMATURES OF ELECTROMAGNETIC SOUND REPRODUCERS—C. R. ROWE and W. H. GERNS, of East Orange, N. J. Filed Aug. 7, 1925, issued Nov. 15, 1927. Assigned to Brandes Laboratories, Inc.

Des. 73,878—RADIO CABINET—DAVID S. SPECTOR, New York, N. Y. Filed June 22, 1927, issued Nov. 15, 1927. Assigned to Federal Telegraph Co.

Des. 73,879—RADIO CABINET—DAVID S. SPECTOR, New York, N. Y. Filed June 22, 1927, issued Nov. 15, 1927. Assigned to Federal Telegraph Co.

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Transferred to the Member grade

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New York	Ithaca	Ballard Wm. C.
	New York City	Lush, W. G.
Ohio	Cincinnati	Jarvis, K. W.
England	London	Strong, Chas. E.

Elected to the Member grade

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	London	Glover, Philip F.
France	Paris	Roosevelt, H. L.

Elected to the Associate grade

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	Stockton, Box 730	Harvey, V. L.
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	Denver, 1538 Gilpin Street	Painter, Chas, E.
	Denver, Western Union Tel. Co.	Rowe, F. S.
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Dist. Columbia	Washington. 4828 Sixteenth St. N. W.	Bliss, A. O., Jr.
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	Chicago, 3982 Barry Avenue	Fast, John E.
	Chicago, Randolph & Canal Sts. c.o Butler	Kamin, Vernon
	Chicago, 9402 Ewing Ave., RCA Radio WGO.	Newell, L. T.
	Galesbury, 1347 W. Main Street	Bellinger, C. E.
	Geneva, Riverbank Laboratories	Crosier, Wm. D.
	Streator, 115 S Vermillion St.	Williams, N. B.
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	Mattapan, 160 Brook Road	Weber, H. C.
	Newton, 13 Newtonville Avenue	Wilson, T. C.
	No. Attleboro, 9 Elm St.	Langlais, G.
	Siasconset	Holden, H. H.
	Sommerville, 22 Oxford St.	Cook, W. H.
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	Detroit, 5479 Stanton Avenue	Manning, S. R.
	Mt. Pleasant, 518 So. Kinney Ave.	Wilbur, D. A.
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	Brooklyn, 65 Pineapple St.	Milania, Clarence
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	Buffalo, 56 Hedley Place	Butler, S. J.
	Buffalo, 158 Box Avenue	Fox, R. A.
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	Freeport, L. I., 209 N. Long Beach Ave	O'Neill, John
	New York City, 602 W. 139th St	Glenn, P. R.
	New York City, 1079 Washington Ave	Hills, B. E.
	New York City, 370 Seventh Ave	Nelson, J. R.
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	New York City, 30 Broad Street	Truax, John
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	Scotia, R. F. D. No. 2	Baker, J. O.
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	Fukushimaken, Haranomachi, Nippon Musen	Ono, Shoji
	Fukushimaken, Haranomachi, Nippon Musen	Sakai, Sadao
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	Tokyo, Nippon Musen Denshin Kaisha	Kobayachi, Riuji
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	Auckland, P. O. Box 779	Sangster, R. L.
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South Africa	Johannesburg, Box 4576	Myers, H.
Wales	Montgomeryshire, Welshpool, Groespluan	Bond, A. E.

Elected to Junior grade

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of
The Institute of Radio Engineers

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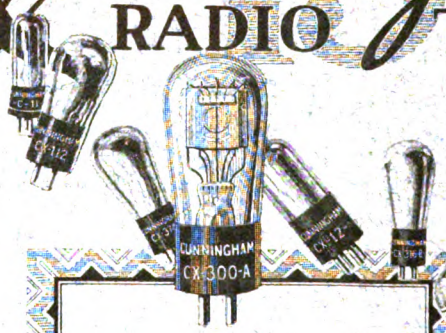

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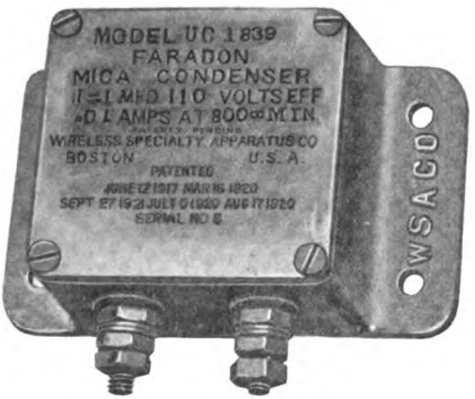

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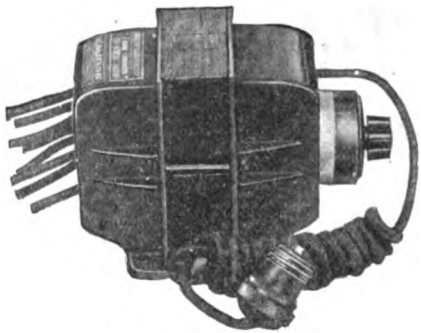

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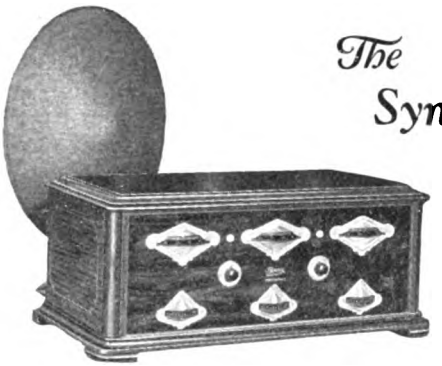

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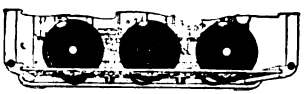

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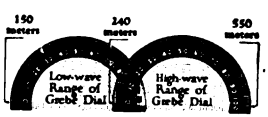

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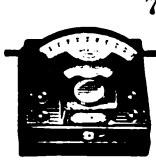

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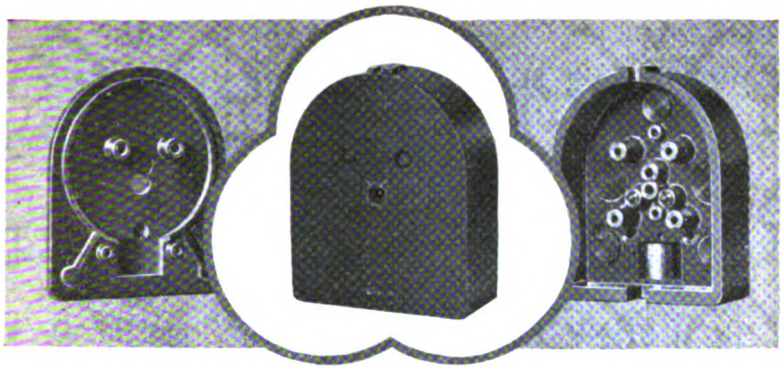

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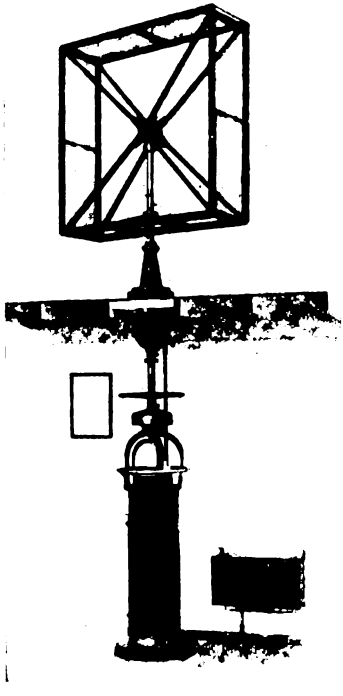

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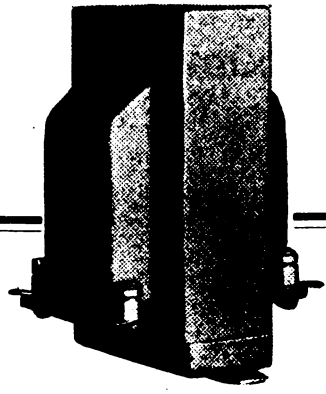

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EXTREMELY successful transformers developed by the Pacent Engineers, who are primarily specialists in Audio Frequency parts. Handsomely finished in dull bronze in two types.

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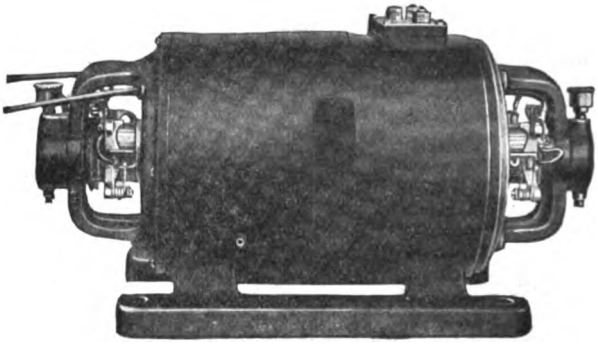

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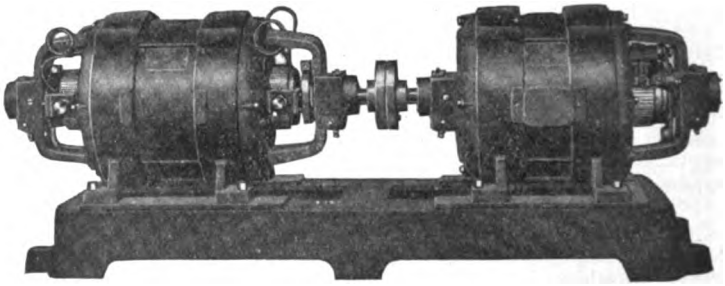

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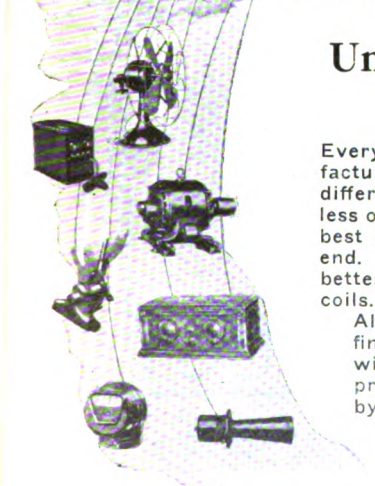

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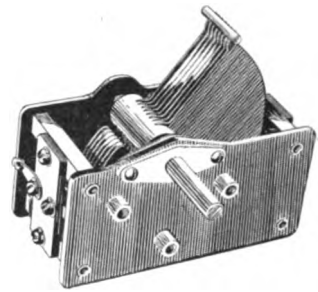

Wizardry and black art harnessed and put to work for entertainment in the home. The Radio. In the expert design and construction of the delicate parts which permit the "talking air" to carry its magic nightly to millions, Scovill takes first rank. Here radio manufacturers find a scientific understanding of their problems and a large scale production which they can convert into additional dollars of profit.

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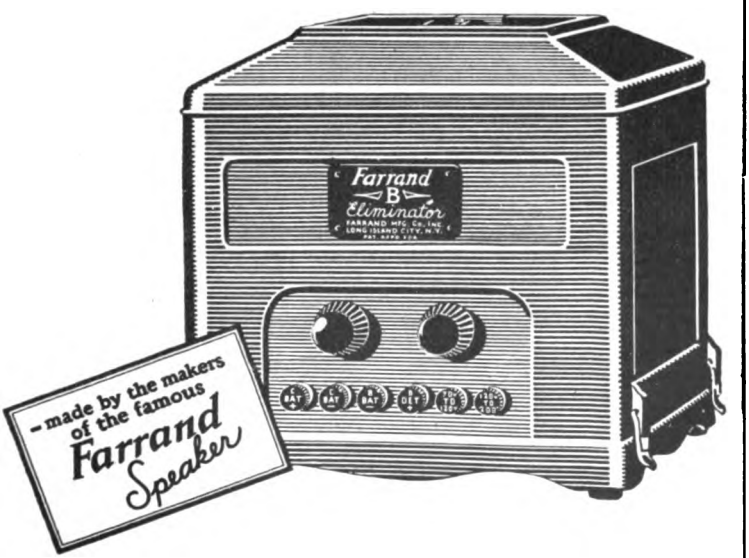

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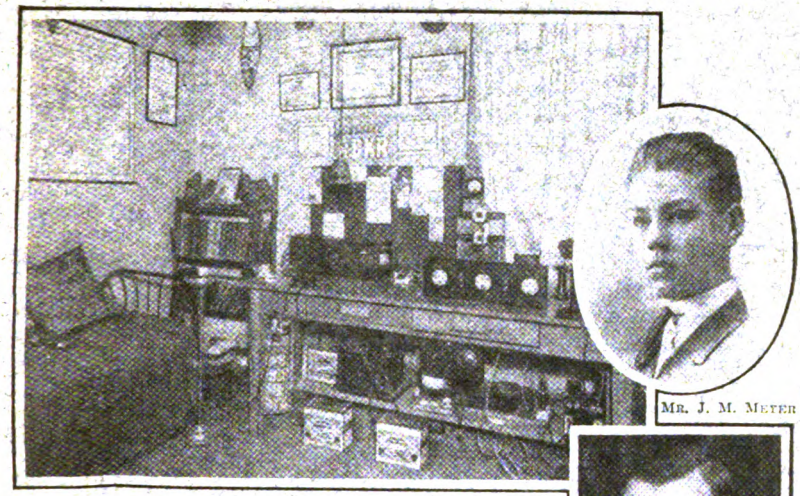

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Milwaukee Radio
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9BKR at Milwaukee Show used Eveready Layerbilt "B" Batteries

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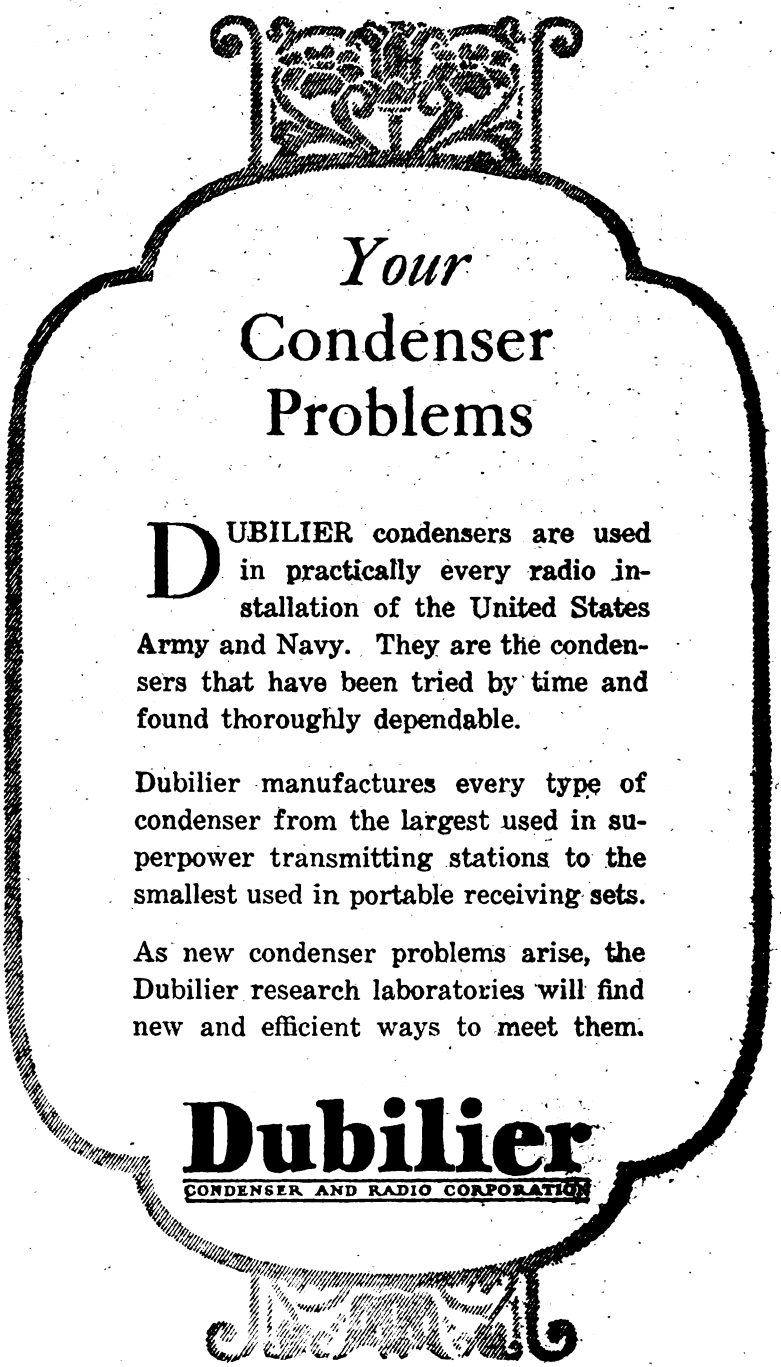

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VOLUME 15 DECEMBER, 1927 NUMBER 12

PROCEEDINGS

of

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Editorial and Advertising Departments
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Radio sets and radio equipment make immensely popular Christmas gifts.

To assure the utmost in Yule-tide Radio enchantment, see that Cunningham Radio Tubes are on duty in every socket. Every owner of a radio will be delighted to receive a set of new tubes.

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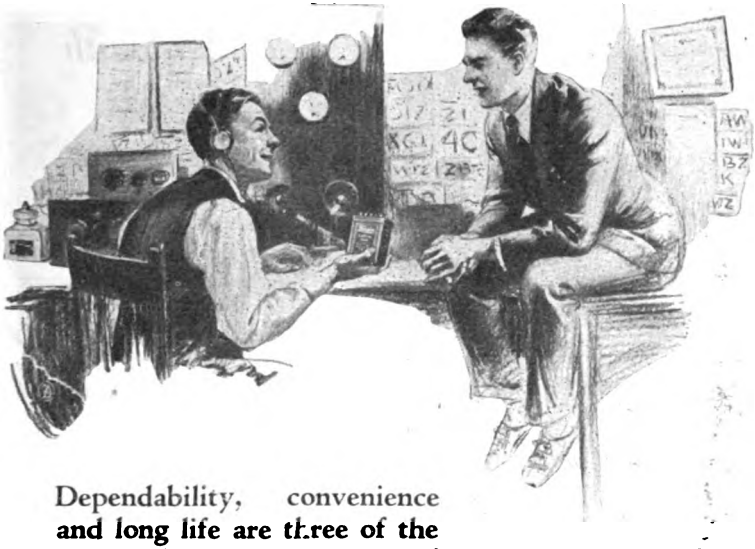

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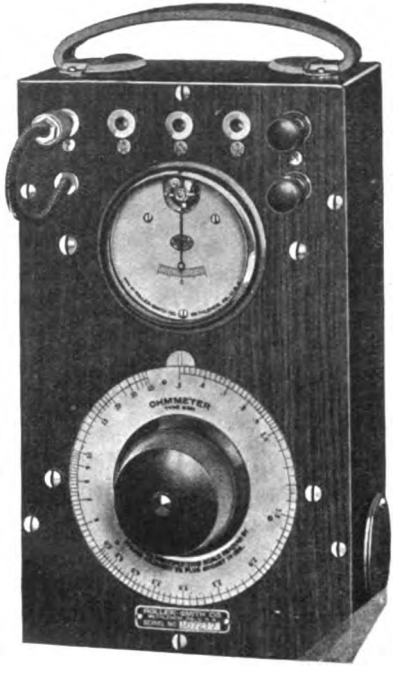

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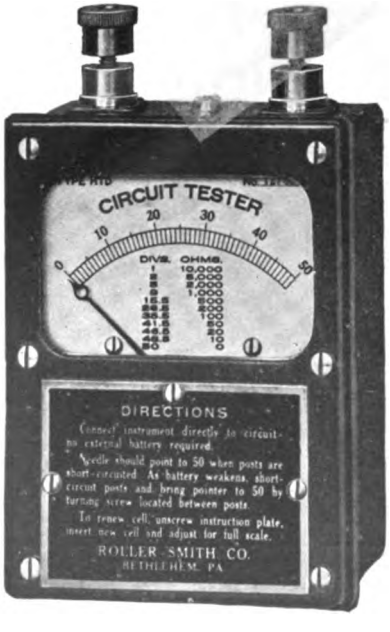
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DIRECTIONS
Connect instrument directly to circuit—no external battery required.
Needle should point to 50 when posts are short-circuited. As battery weakens, short-circuit posts and bring pointer to 50 by turning screw located between posts.
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ROLLER-SMITH CO.
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A Complete
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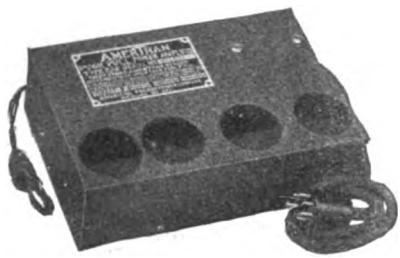

THE *new* AMERTRAN PUSH-PULL Power Amplifier is a completely assembled two-stage unit embodying a first stage AMERTRAN DeLuxe followed by AMERTRAN Input and Output Transformers for Power Tubes. The metal box has four sockets, two for power tubes, and a four prong and a five prong socket in the first stage for either a standard amplifying tube or the UX-201 A type, or a UY-227 (C-327) A.C. tube. Using the latter tube, the amplifier can be entirely A.C. operated.

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Send for further information. Inquiries on this new unit will be handled promptly.

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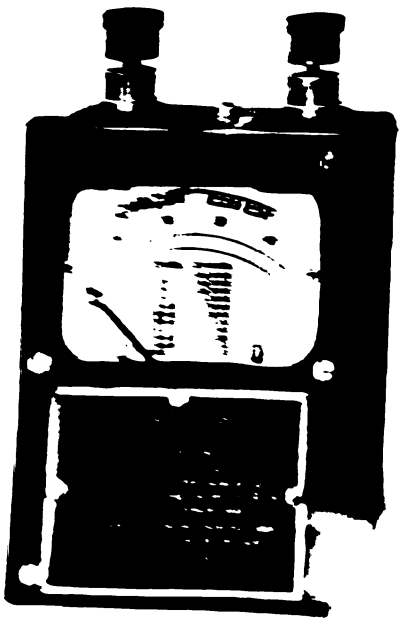

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Diameter 2 inches

Pattern No. 190 A.C. Voltmeter

The advent of vacuum tubes having filaments adaptable for excitation from alternating current creates the necessity for an absolute means of control other than the filament rheostat, for it is frequently found that a particular setting of the filament voltage is necessary to eliminate objectional hum.

For A.C. filament control the new Jewell A.C. voltmeter, Pattern No. 190 is very desirable. Jewell Pattern No. 190 instruments have a body diameter of 2 inches and in general appearance are similar to the Jewell Pattern No. 135 direct current style. The movement is of the moving vane type with special modifications for the small size case. The instruments are accurate and neat in appearance and are designed for continuous service. Energy consumption is very small.

Pattern No. 190 instruments are available in many ranges, of which the 0-1.5, 0-3, 0-15 and 0-150 volt ranges are most in demand. The 0-1.5 volt range is suitable for tubes having a direct heated Cathode, the 0-3 volt range for tubes having an indirectly heated Cathode, and the 0-150 volt range where primary control is used.

Write for a copy of our 15-C catalog supplement which describes this instrument in detail.

Jewell Electrical Instrument Co.

1650 Walnut St. - Chicago

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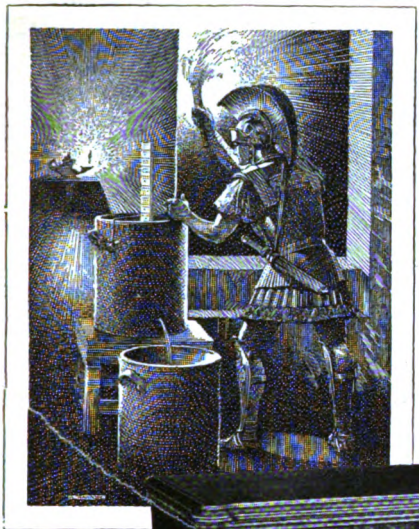

About 300 B. C. the Greeks sent messages by means of the Clepsydra. In two stations were water jars of equal size in which floated marked rods. On signal by torch or flag, plugs were pulled. When the rods lowered to a certain mark, at another signal both plugs were reinserted. The rods in both towers dropping to the same mark, the message was known.

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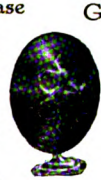

Grebe construction, backed by eighteen years of making nothing but the highest grade radio apparatus, guarantees the durability of these superior qualities.

Grebe Synchrophase Seven, $135; Grebe Natural Speaker, $35. Send for Booklet I; then ask your dealer to demonstrate, in your home, that you can "get it *better* with a Grebe."

GREBE SYNCHROPHASE
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RADIO

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Makers of quality radio since 1909

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You can candle

an *egg*—but not a *battery*

THERE isn't much difference in the size or shape of batteries. And you can't tell how good they are before you use them. ¶ If you could, one element alone would win your preference for Burgess. That element is *Chrome*. *Chrome* is the *preservative* that maintains an abundance of unfailing energy in Burgess Batteries—long after most dry cells cease to function. The black and white stripes are individual marks for identifying Burgess *Chrome* Batteries. Buy them for long lasting, dependable performance!

Chrome—the preserving element used in leather, metals, paints and other materials subject to wear, is also used in Burgess Batteries. It gives them unusual *staying* power. Burgess *Chrome* Batteries are patented.

Ask Any Radio Engineer

BURGESS BATTERY COMPANY

General Sales Office: CHICAGO

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BURGESS BATTERIES

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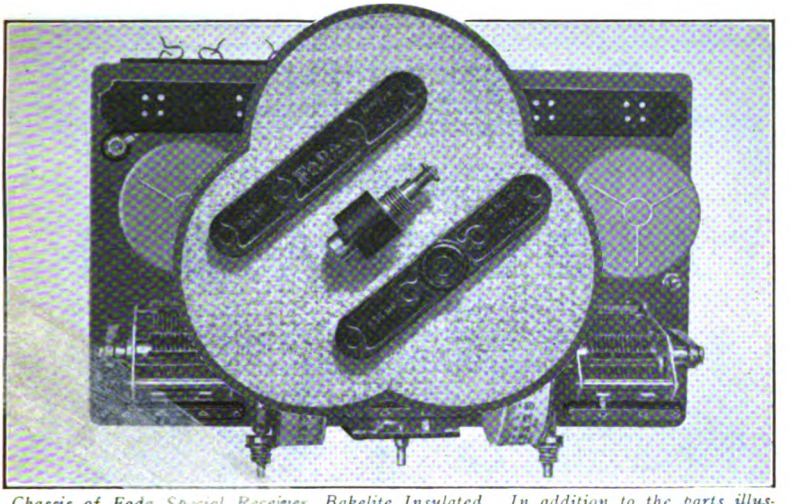

Chassis of Fada Special Receiver, Bakelite Insulated. In addition to the parts illustrated, Bakelite Materials are also used for socket lower plate R.F. transformer, condenser and leak mounting strips and minimum adjustor base. This Receiver is made by F. A. D. Andrea, Inc., Long Island, N.Y.

In the Fada Special

vital parts are insulated with Bakelite Materials

RECOGNIZING that sustained excellence of performance may only be assured through the use of an insulation whose properties are unimpaired by time, exposure or use, F. A. D. Andrea, Inc., have adopted Bakelite Materials for many parts of the Fada Special Receiver.

The use of Bakelite Materials not only provides excellent insulation, but economical production as well. For example, twenty of the upper and lower condenser strips are molded at one time with holes, recesses, and relief lettering accurately formed. In the case of the neutrodon cylinder part a nine cavity mold is used, the Bakelite Molded cylinders being formed around the terminal posts.

Since the earliest days of Radio, the Bakelite Corporation has kept pace with the development of Broadcast Receivers, and its engineers and research laboratories have co-operated with manufacturers in adapting Bakelite Materials to each exacting need. The wealth of knowledge acquired during these years of experience is offered to the radio industry through our Service Engineers located in important industrial centers.

Write for Booklet 33

BAKELITE CORPORATION

247 Park Avenue, New York, N.Y. Chicago Office, 635 W. 22nd Street
BAKELITE CORP. OF CANADA, LTD. 163 Dufferin Street, Toronto, Ont.

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RADIOTRONS
for modern
A. C. OPERATION

POWER AMPLIFIER RADIOTRON UX-171

Filament—5 Volts—.5 Amperes

Plate Voltage	90	135	180	Volts
Negative Grid Bias	16½	27	40½	Volts
Plate Current	10	16	20	Milliamperes
Plate Resistance (A.C.)	2500	2200	2000	Ohms
Mutual Conductance	1200	1360	1500	Micromhos
Voltage Amplification Factor	3.0	3.0	3.0	
Max. Undistorted Output	130	330	700	Milliwatts

R. F. & A. F. AMPLIFIER RADIOTRON UX-226

Filament {A. C.} 1.5 Volts—1.05 Amperes

Plate Voltage	90	135	135	180	Volts
Negative Grid Bias	6	12	9	13½	Volts
Plate Current	3.7	3	6	7.5	Milliamperes
Plate Resistance (A.C.)	9400	10,000	7400	7000	Ohms
Mutual Conductance	875	820	1100	1170	Micromhos
Voltage Amplification Factor	8.2	8.2	8.2	8.2	
Max. Undistorted Output	20	60	70	120	Milliwatts

DETECTOR RADIOTRON UY-227

Heater {A. C.} 2.5 Volts—1.75 Amperes

Plate Voltage	45	90	Volts
Grid Leak	2-9	¼-1	Megohms
Plate Current	2	7	Milliamperes
Plate Resistance (A.C.)	10,000	8000	Ohms
Mutual Conductance	800	1000	Micromhos
Voltage Amplification Factor	8	8	

FULL WAVE RECTIFIER RADIOTRON UX-280

A.C. Filament Voltage	5.0	Volts
A.C. Filament Current	2.0	Amperes
A.C. Plate Voltage (Max. per plate)	300	Volts
D.C. Output Current (Maximum)	125	Milliamperes
Effective D.C. Output Voltage of typical Rectifier Circuit at full output current as applied to Filter	260	Volts

HALF WAVE RECTIFIER RADIOTRON UX-281

A.C. Filament Voltage	7.5	Volts
A.C. Filament Current	1.25	Amperes
A.C. Plate Voltage (Max. per plate)	750	Volts
D.C. Output Current (Maximum)	110	Milliamperes
Effective D.C. Output Voltage of typical Rectifier Circuit at full output current as applied to Filter	620	Volts

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New York Chicago San Francisco

RCA Radiotron

MADE BY THE MAKERS OF THE RADIOLA

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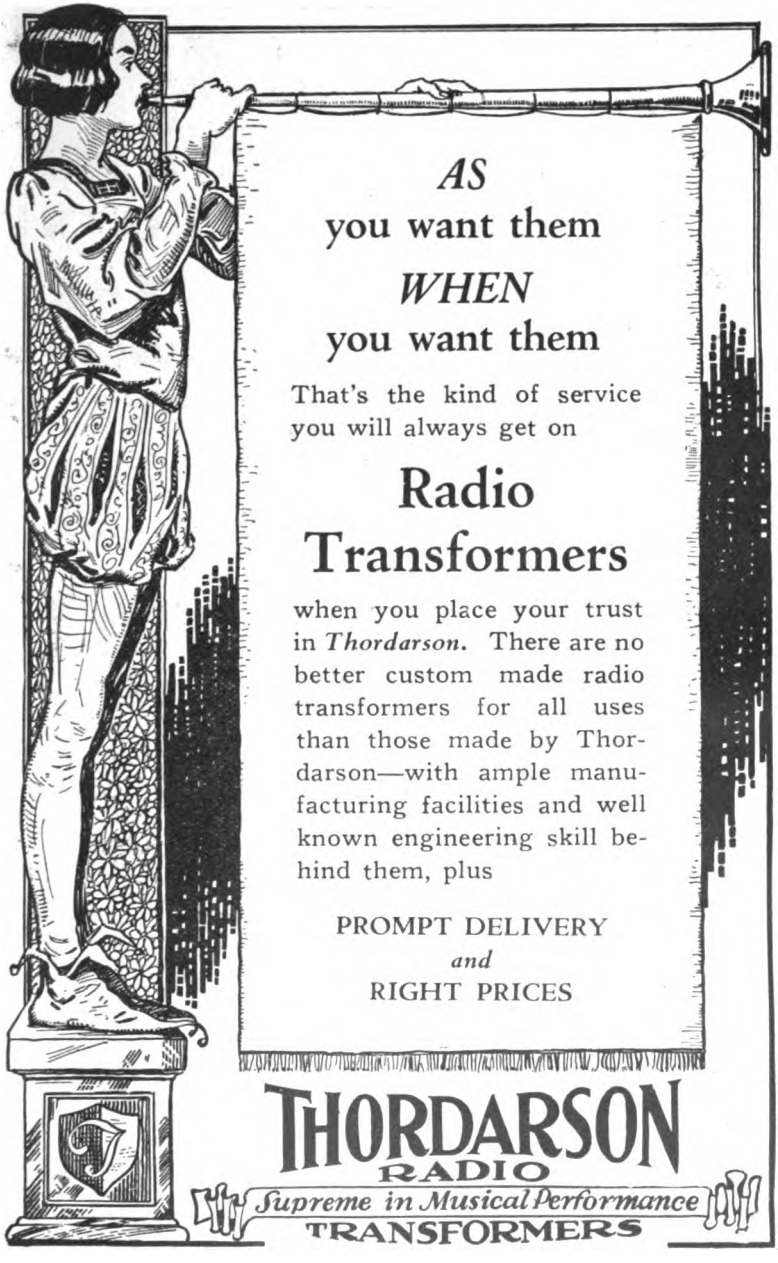
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WHEN
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That's the kind of service you will always get on
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and
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Supreme in Musical Performance
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Transformer specialists since 1895

WORLD'S OLDEST AND LARGEST EXCLUSIVE TRANSFORMER MAKERS

Chicago, U.S.A. (3570)

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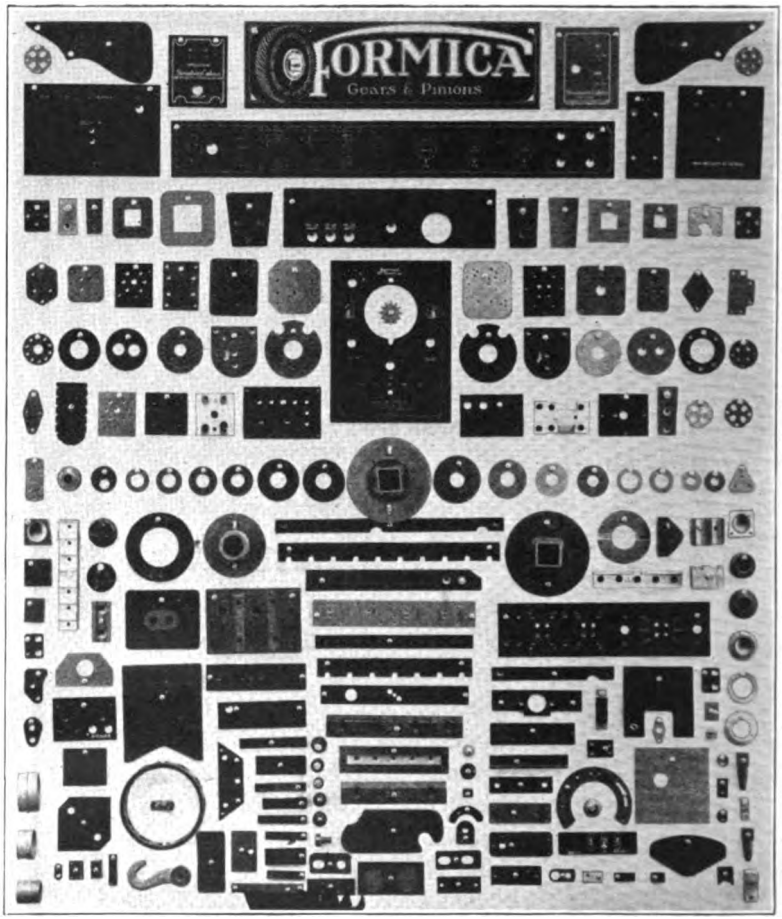

Formica Parts for Many Purposes

Formica is provided for all sorts of radio electrical and mechanical purposes in a great variety of shapes and sizes—punched, drilled and machined in a variety of ways. It combines excellent quality as an insulator for high or low frequency work, with mechanical strength, immunity to moisture, absorption and chemical inertness.

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Made from Anhydrous Bakelite Resins

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Formica has a Complete Service on Insulating Material for Radio Manufacturers

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A New Standard of A-C Operation With Arcturus A-C Tubes

The unique advantages which we claim for Arcturus A-C Tubes are directly traceable to unique features of construction and exceptional operating characteristics.

The exceptionally long life of Arcturus Tubes is due to the enormous electron supply, resulting from the heater operating at a low temperature.

The highly efficient cathode is responsible for the unusual sensitivity of Arcturus A-C. Tubes and for the exceptional volume and tone quality which their use insures. This cathode produces: 1. A high amplification factor (10.5). 2. A low plate impedence (9,000 ohms). 3. A high mutual conductance (1160 micromhos).

Since the base of the Arcturus A-C Tube is of the standard four-prong type, no additional terminals are required, making Arcturus Tubes adaptable to existing circuits with all the simplicity of D-C tubes. No center taps or balancing are required. A common toy transformer may be used. Filament voltage is the same (15 volts) for all types, detector, amplifier and power.

The freedom from hum which is one of the most important features of Arcturus A-C Tubes, is due to the use of low A-C current, only 0.35 ampere. Arcturus Tubes in all stages are four element tubes with indirectly heated cathodes.

Normal variations in line voltage do not affect the operation of Arcturus A-C Tubes. The amplification factor is practically constant over a wide range of filament voltages—13.0 to 18.0 volts.

ARCTURUS RADIO COMPANY, INC.
255 Sherman Avenue, Newark, N. J.

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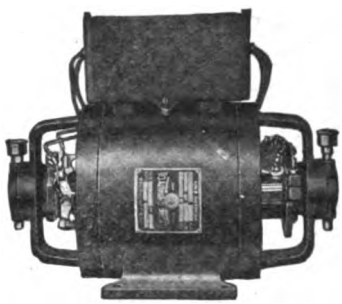
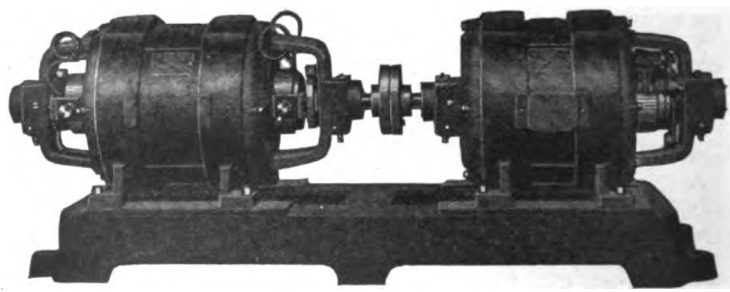
A.C. radio receivers in the D.C. Districts? *Certainly!* with an "ESCO" dynamotor or motor-generator —



Dynamotor with Filter for Radio Receivers

The type R143 rotary converters are suitable for phonographs with vacuum tube amplifiers such as the Orthophonic Victrolas and Brunswick Panatropes without radio receivers. No filters are required.

The dynamotors and motor generators are suitable for radio receivers and for combination instruments containing phonographs and receivers. Filters are usually required. The dynamotors and motor-generators with filters give as good or better results than are obtained from ordinary 60-cycle lighting sockets. They are furnished completely assembled and connected and are very easily installed.



"ESCO" two and three unit sets have become the accepted standards for transmission. The "ESCO" line consists of over 700 combinations. These are covered by Bulletin 237D.

Our engineers are always willing to cooperate in the development of special sets.

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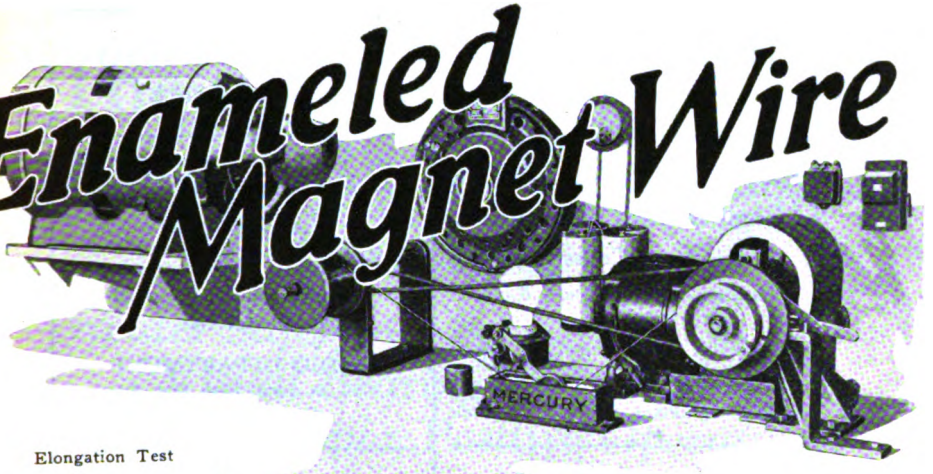
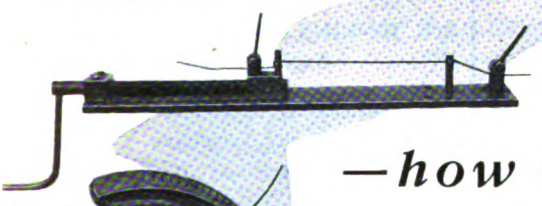
Enameled Magnet Wire

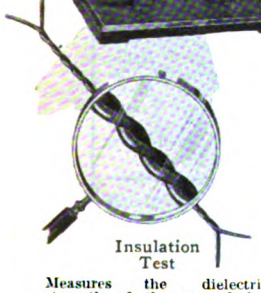


Elongation Test



Mercury Bath Test
A positive method of detecting and guarding against imperfections in the enamel insulation.

Reactance Test
Determines the softness of the wire. Checks the completeness and uniformity of the annealing.



Insulation Test
Measures the dielectric strength of the enamel insulation by determining the voltage required to break it down.

—how would YOU test it?

Regardless of how varied or severe your tests might be, Dudlo enameled wire has already been subjected to a series of scientific tests at the factory, covering every physical or electrical characteristic required for any purpose.

Besides the four tests illustrated here, a number of others regularly employed including the "Ageing" test to determine the ability of the enamel to stand up under heat over continued use—also the "Insolubility" test proving the qualities of the insulation to resist the effects of oil, water, varnish, etc.

Thus Dudlo protects the quality of *your products* by constantly maintaining the highest possible standard of its own.

DUDLO

DUDLO MANUFACTURING CORPORATION, FORT WAYNE, INDIANA

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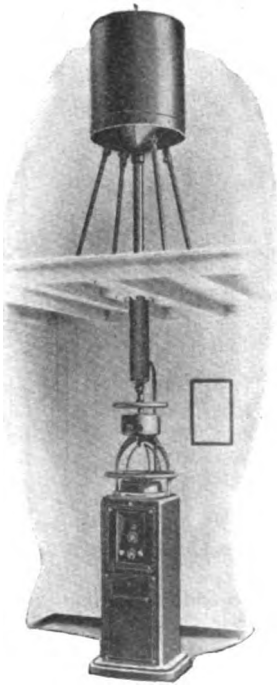
The new Kolster Radio Compass provides

Greater Safety— Visual Bearings— Simple, Positive Operation



From its aluminum pedestal to the new enclosed loop, the new Kolster Radio Compass [Type AM-4490] embodies every improvement radio science has to offer. The insulated cylindrical housing of the loop affords complete protection against wind, ice, snow, and spray. A tiny lamp flashes the signals of nearby stations. While for long-distance bearings, a Kolster eight-tube receiver with a new circuit especially designed for radio compass work provides the maximum of selectivity and sensitiveness.

No knowledge of radio is necessary to operate the Kolster Radio Compass. It is built for the navigator. The positive unidirectional indicator gives instantaneous readings without guesswork or computation. A new and improved Kolster automatic compensator corrects all natural errors, and makes *direct* readings absolutely accurate.

Kolster Radio is also setting new standards of performance in brose

FEDERAL-BRANDES, INC.

200 Mt. Pleasant Avenue
Newark, N. J.

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In the Still of the Arctic Night

IN THE STILL of the Arctic night, broken only by the occasional bark of a far-off walrus, the Radio Operator cannot leave his ship, frozen into the ice pack, and run around the corner to the Radio Shack for a new filter condenser to replace the one that just blew out.

He has got to be sure when he starts that his equipment is not going to give out.

That is why Cliff Himoe took TOBE Condensers with him on the *Bowdoin* for the MacMillan Arctic Expedition.

Here is a Radio message just received from the boat: The TOBES are up to the mark!

"YOUR FILTER CONDENSER STANDING UP WELL ON BOWDOIN'S TRANSMITTER, WITH NO SIGNS OF TROUBLE AT CONTINUOUS 2000 VOLTS D.C. REGARDS FROM THE ARCTIC."

HIMOE WNP

The University-of-Michigan-Greenland Expedition is also equipped with TOBE Condensers.

Make sure that your Radio Power Equipment includes TOBE Condensers. TOBE Condensers cure condenser worries permanently and painlessly.

Tobe Deutschmann Co.

Cambridge :: Mass.

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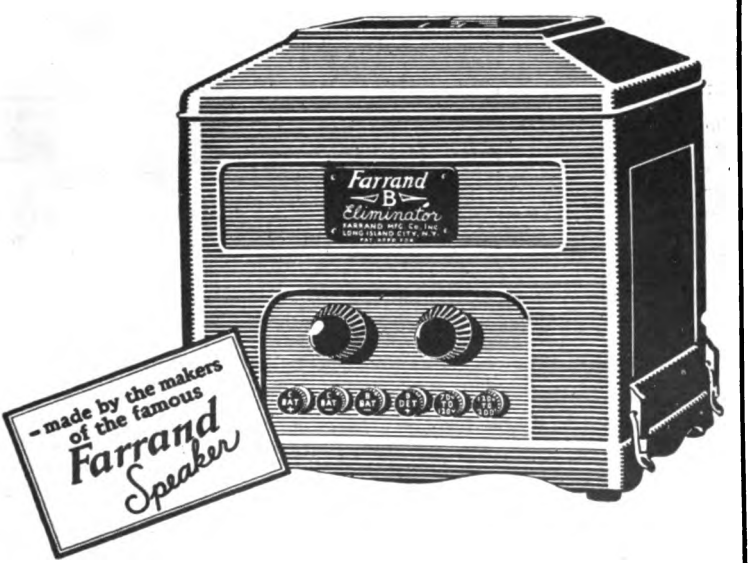
Farrand
B
ELIMINATOR



KNOWN QUALITY

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for better Radio Reception

Irrespective of the receiving circuit employed, it can be materially improved by micrometric resistance, or simply specifying

VOLUME CONTROL CLAROSTAT

This device is compact, good looking, inexpensive and handy. It is becoming most popular in manufactured, custom-built and home-made radio sets as a volume control and as a plate voltage control for r.f. and detector tubes. It may be applied as a loud-speaker control. It serves to match transformers for best tone quality. It can be used as a tone control in conjunction with a fixed condenser. It is an ideal micrometric control of regeneration particularly on short waves.

The Volume Control Clarostat has a resistance range of practically zero to 500,000 ohms in several turns of the knob. It has ample current-carrying capacity for receiver applications. It is silent in operation. One-hole mounting. Convenient terminals. Absolutely dependable. *And only $1.50.*

Of course you must continue to specify the Standard Clarostat for B-eliminator applications, the Heavy-Duty Clarostat for line-voltage control, and the Power Clarostat for heavy-duty B-power units, A-B-C radio power units, electrified sets, and other applications.

There's a Clarostat for every purpose. Make sure, however, you specify and get genuine Clarostats, with the name CLAROSTAT stamped on the shell. Don't jeopardize your engineering and production by utilizing questionable resistors!

Write us for data on all types and ranges of Clarostats, and place your resistance problems up to us.

AMERICAN MECHANICAL LABORATORIES, INC.

MEMBER R.M.A.

Specialists in Variable Resistors

285 N. Sixth St., BROOKLYN, N. Y.

—CLAROSTAT—

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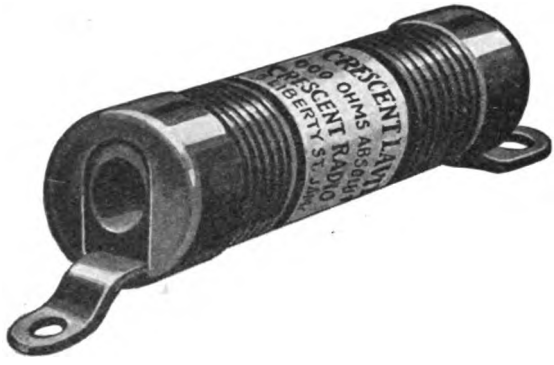

Crescent Lavite

is the pioneer resistance and still in the van. Economical in size but gluttons for punishment

Used by U. S. Government, General Electric Co., Bell Telephone Laboratories, Westinghouse Electric and Manufacturing Co., and ninety broadcast stations.

Write for data which will be supplied on request.

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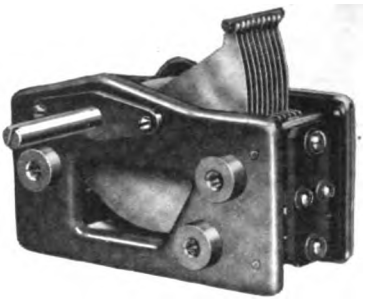

Hidden parts of delicate construction—heart of the radio. How truly present-day "music of the spheres" depends on mechanical exactness for proper reception! Scovill accuracy in fabrication and capacity for large-scale production bring radio manufacturers to us as a matter of course.

A SCOVILL PRODUCT

Scovill is the name of a broad service to industry. It places acres of factories, forests of machinery, hosts of skilled workmen, metallurgists, modern laboratories and trained representatives at the disposal of those who require parts or finished products of metal. Why not see how Scovill can serve you? Call our nearest office.

THESE ARE SCOVILL PRODUCTS

Made to Order: Condensers, Variable and Variable Vernier and parts for same, Metal Stampings, Screw Machine Products, Switches, Decorated Metal Radio Panels, Parts from Brass, Steel and Aluminum. *Carried in Stock:* Butts and Hinges, Continuous Hinges, Machine Screws. *Brass Mill Products:* Sheet, rod, wire, tubing.

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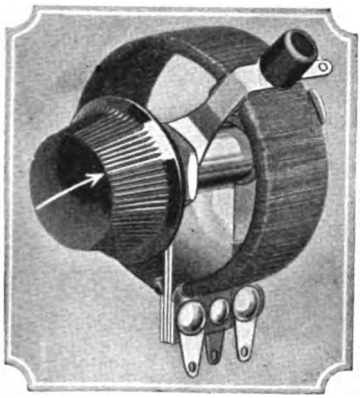

Greater Efficiency and Economy in

"B" Eliminator Resistances

THE new Centralab 4th Terminal Potentiometers have two variable contact arms on each unit. Two of these units in series will provide complete voltage regulation for any "B" power supply without additional fixed resistors or variables. One additional unit will provide two "C" bias taps when desired. The economy is apparent in that there are fewer units to buy, and less assembly time to mount them on the panel.

Centralab 4th Terminal Potentiometers are wire wound on a frame of metal and asbestos. They will safely dissipate in excess of 30 watts without break down. This high current carrying capacity makes possible a low total resistance across the "B" supply, giving much better voltage regulation than the high resistances normally used, and sufficient current load on open circuit to substantially lessen the danger of condenser break down.

Fourth Terminal Potentiometers are wire wound in resistance values up to 6000 ohms. The diameter is 2", depth ¾". They are recommended as the best and most economical of available "B" power voltage controls.

Where smaller units must be used because of small panel space, there are other Centralab wire wound potentiometers with diameters of 1⅝" and 1⅜" respectively that can be furnished in resistances up to 20,000 ohms, and variable high resistances up to 500,000 ohms.

Complete information and circuit data will be gladly mailed to those interested.

Central Radio Laboratories
16 Keefe Avenue, Milwaukee, Wisconsin

Centralab

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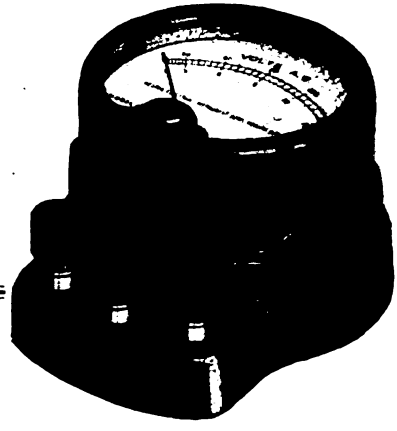

New Instruments for A. C. Operation

TO meet the new need, Weston has developed the Model 528 A.C. instruments—small, compact and portable—for testing A.C. operated radio sets.

These instruments represent a distinct achievement in high-grade, small A.C. meter design and construction. They are responsive and excellently damped, may be used continuously at full scale reading, have open scales and are accurate on any commercial frequency. They are made as Voltmeters and Ammeters, the former having an exceptionally high internal resistance, self-contained in double ranges up to 600 volts.

So that they may be readily distinguished from Model 489, their companion D.C. instruments, these new A.C. Portables have red and black mottled bakelite cases instead of the plain black.

Circular J, a very complete little catalogue covering the entire line of Weston Radio Instruments, will be mailed upon request. Everyone interested in radio will find it a valuable reference.

WESTON ELECTRICAL INSTRUMENT CORPORATION
73 Weston Ave. Newark, N. J.

Weston
PIONEERS SINCE 1888
INSTRUMENTS

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(Fill in and mail to the Institute of Radio Engineers, 37 West 39th Street, New York, N. Y.)

To the Committee on War Memorial for American Engineers:

The undersigned, a member of the Institute of Radio Engineers, encloses herewith a donation to the fund for the creation of a war memorial to American Engineers who gave their lives in service out of this country during the World War.

_____
(Name)

_____
(City)

_____
(State)

The following members of the Institute of Radio Engineers are known to have given their lives in service outside this country during the war:

Name	*Died at*

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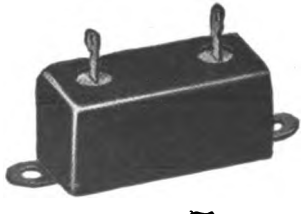

Service *to* Manufacturers

MILLIONS of Fast condensers are in daily use in radio sets made by leading set manufacturers. They offer conclusive proof as to their extraordinarily high insulation resistance and excellent electrical characteristics. Special one-piece die-press steel housing seals and protects them permanently from climatic conditions or abuse. Built for endurance, long life and reliability.

Manufacturers looking for a dependable source of supply will find here one of the largest organizations of its kind in the world. Specialization keeps our costs low and quality high.

Send us your specifications. We make everything in the condenser line.

JOHN E. FAST & CO.

Established 1919

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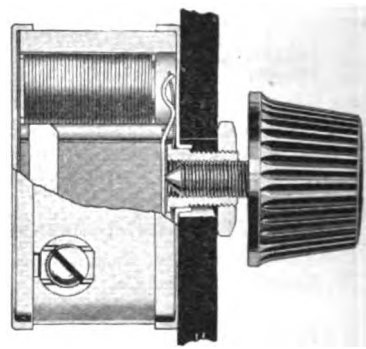

Because of its unusually wide range and accuracy of control, the Bradleyohm-E is extensively used by B-eliminator manufacturers. The resistance of the scientifically-treated discs is not affected by use or atmospheric conditions. It does not vary in service. When once the correct value is selected, it remains fixed.

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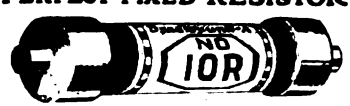

This molded resistor is baked under high pressure, and then accurately calibrated. Bradleyunits are guaranteed to be accurate within 5% of their ratings. Unaffected by temperature, moisture or age.

Bradleyunit-B

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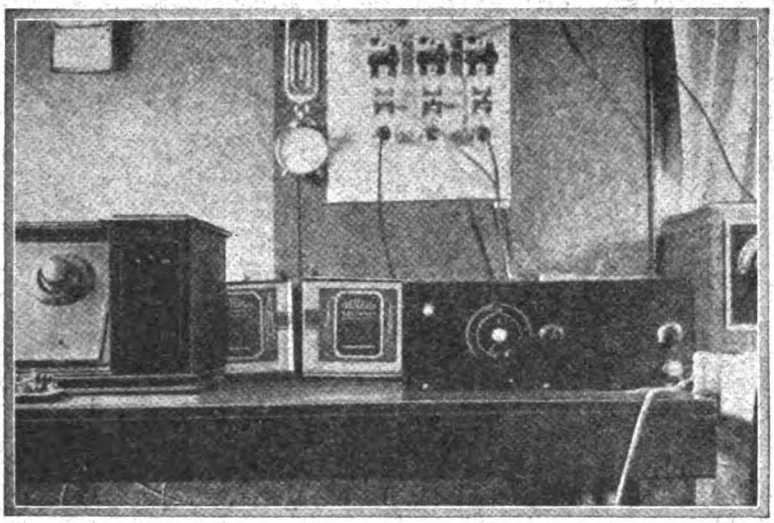

It's not the bottle but the *kind of juice* that counts

USING a UX-199 tube as an oscillator with Eveready No. 770 Heavy-Duty "B" Batteries for plate power, H. A. Mackley, 9 LF, Peoria, Ill., has done some remarkable work. "Daylight communication has been held with the 5th and 6th District, New Mexico and California, on the south and west," he says, "and with the 1st and 3rd on the east." In this and in more severe tests, the 770's "stood up exceptionally well and the curve was very flat the first 200 hours' use. I heartily recommend Eveready batteries, especially the 770 and the Layerbilt, for amateur use on up to 5-watt tubes. If some of the amateurs would try to perfect their apparatus instead of crowding some of the 50-watt bottles to the limit, they would get remarkable distance with a 201-A or a 5-watter." Eveready "B" Batteries produce pure DC only, which is why they are the preferred source of plate power not only for short wave work, but also for BCL receivers.

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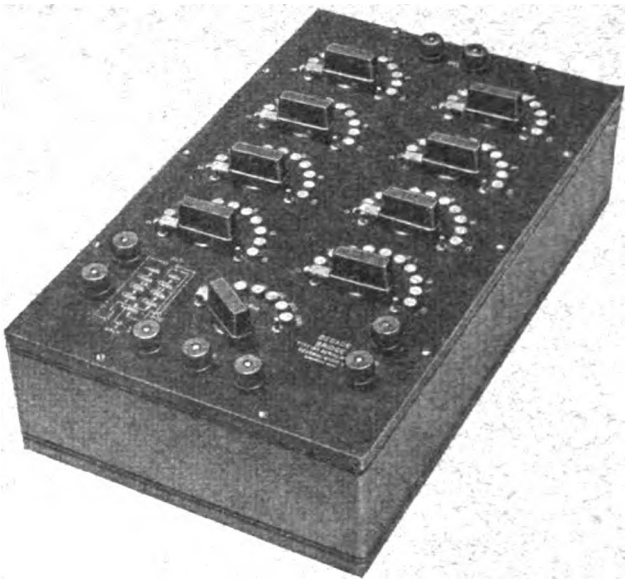

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